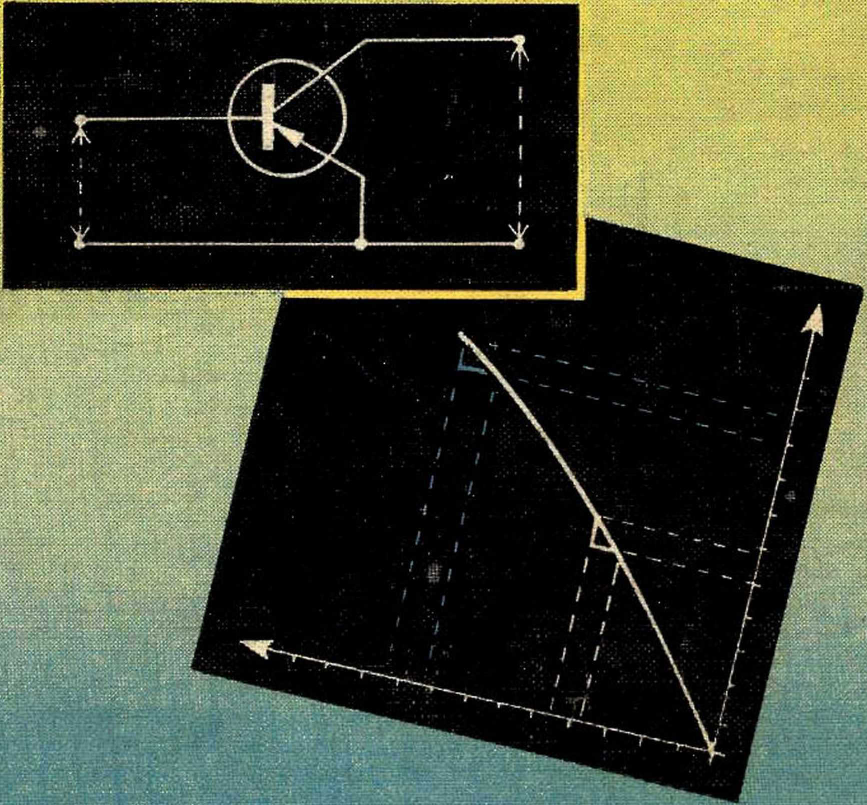
G. FONTAINE



DIODES AND TRANSISTORS

GENERAL PRINCIPLES

PHILIPS TECHNICAL LIBRARY

DIODES AND TRANSISTORS

DIODES AND TRANSISTORS

GENERAL PRINCIPLES

G. FONTAINE

1963

PHILIPS TECHNICAL LIBRARY

Translated from French by P. J. Arthern, Carshalton, England

Publisher's note:
This book contains vi + 470 pages, 445 illustrations
Other editions in Philips Technical Library in the German,
French, Spanish and Dutch language

U.D.C. 621 . 382 . 2/3

Original French edition:
© N.V. Philips' Gloeilampenfabrieken, Eindhoven (The Netherlands), 1961

English edition:
© N.V. Philips' Gloeilampenfabrieken, Eindhoven (The Netherlands), 1963

All rights reserved

No representation or warranty is given that the matter treated in this book is free from patent rights; nothing herein should be interpreted as granting, by implication or otherwise, a licence under any patent rights

Printed in The Netherlands

PREFACE

The expansion of the scope of electronics and its recent introduction into many branches of industry is due, to a large extent, to the advent of the semiconductor. Are the principles of these new elements so different from the equivalents required for the study and setting up of electronic circuits employing vacuum tubes?

Semiconductors do indeed differ from vacuum tubes in principle, but also in the terminology employed to define the various parameters.

Besides, the ability to follow the extremely rapid evolution of this technique demands from the specialist perfect knowledge of the language and a "pattern of thinking" based on the study of physics and electronics with reference to semiconductors.

The interpretation of the terms used does not offer any special problems when dealing with voltage or current but, on the other hand, the use of characteristic curves or the choice of parameters required for the calculation of a stage confronts us with problems that cannot be solved by means of the habitual logic developed in the study of tube circuits.

The aim of this work whose main care has been to avoid mathematical demonstrations, is to illustrate the fundamentals, either by physical explanations or by the very frequent use of graphs, to the extent required for the application of semiconductors in the most favourable conditions.

June, 1963 G. FONTAINE

CONTENTS

PART ONE – PHYSICAL PHENOMENA IN SEMICONDUCTORS

CHAPTER 1 – GENERAL CONSIDERATIONS 8
1.1. Characteristic properties of semiconductors 8

CHAPTER 2 – DEFINITION. 12
2.1. Structure 12
2.2. Bonds between two germanium atoms 12
2.3. Molecular structure 15
2.4. Ways in which equilibrium can be disturbed 15
2.5. Increase of temperature 15
2.6. Conduction by holes 16
2.7. Conductivity in a block of germanium 16
2.8. *N*-type germanium 19
2.9. *P*-type germanium 23
2.10. Representation of a block of pure germanium 24
2.11. Conduction in a block of *N*-type germanium 24
2.12. Conduction in a block of *P*-type germanium 27
2.13. Conduction in a block of pure germanium 31
2.14. The *PN*- junction 31
2.15. The junction capacitance 36
2.16. The effect of an external electric field on the junction 36

PART TWO – DIODES

CHAPTER 3 – JUNCTION DIODES 42
3.1. Principal characteristics 42

CHAPTER 4 – POINT-CONTACT DIODES 46

CHAPTER 5 – ANALYSIS OF DIODE CHARACTERISTICS 49
5.1. Diode biased in the forward direction 49
5.2. Diode biased in the reverse direction 53

CHAPTER 6 – THE BREAKDOWN OF A JUNCTION 57
6.1. Thermal breakdown 57
6.2. Avalanche breakdown 58
6.3. Zener breakdown 58

CHAPTER 7 – THE INFLUENCE OF TEMPERATURE 61
7.1. The influence of temperature on the reverse characteristic 61
7.2. The influence of temperature on the forward characteristic 62
7.3. Explanation of the difference in the influence of temperature on the forward and reverse characteristic of a diode 65

7.4. Diode biased in the forward direction 66
7.5. Diode biased in the reverse direction 66

CHAPTER 8 – COMPARISON BETWEEN THERMIONIC DIODES AND SEMICONDUCTOR DIODES 69

CHAPTER 9 – RECTIFICATION. 70
9.1. Maximum permissible reverse voltage $-V_{DM}$ 70
9.2. The value of the load resistance 74
9.3. Non-sinusoidal input voltage 74
9.4. The rectified current 74

CHAPTER 10 – COMPARISON BETWEEN THE OPERATION OF A SEMICONDUCTOR DIODE AND A THERMIONIC DIODE AS DETECTOR. 77
10.1. Detection efficiency 77
10.2. Determination of the optimum value of the load resistance for a semiconductor diode and for a thermionic diode 81
10.3. Low load resistance *R* 81
10.4. High load resistance *R*. 85
10.5. The effect of temperature on the detection efficiency of a germanium diode 86
10.6. Damping resistance 86

CHAPTER 11 – OPERATION AT RADIO FREQUENCIES 90

PART THREE – TRANSISTORS

CHAPTER 12 – GENERAL CONSIDERATIONS 94

CHAPTER 13 – THE JUNCTION TRANSISTOR 97
13.1. Circuit symbols 97
13.2. Conduction in a transistor. 97

CHAPTER 14 – THE TECHNOLOGY OF THE TRANSISTOR 98
14.1. Pure germanium or silicon 98
14.2. The manufacture of *N*-type germanium 101
14.3. The manufacture of junction transistors 101

CHAPTER 15 – THE MODE OF OPERATION OF THERMIONIC VALVES 102
15.1. Advantages 102
15.2. Limitations 102

CHAPTER 16 – THE MODE OF OPERATION OF TRANSISTORS 105

16.1. Advantages 105
16.2. Limitations 105

CHAPTER 17 – *PNP* TRANSISTORS 106

17.1. Formation of the *PN*-junction between regions *A* and *B* 106
17.2. Formation of the *NP*-junction between regions *B* and *C* 109
17.3. The mode of operation of a *PNP* transistor 110
17.4. Reversal of the voltage V_C 113
17.5. Reversal of the voltages V_C and V_B 114
17.6. Leakage current 117

CHAPTER 18 – *NPN* TRANSISTORS 118

18.1. Formation of the *NP*-junction between regions *A* and *B* 118
18.2. Formation of the *PN*-junction between regions *B* and *C* 121
18.3. The mode of operation of an *NPN* transistor 122
18.4. The construction of a junction transistor 125

CHAPTER 19 – CONSIDERATION OF TRANSISTOR CHARACTERISTICS 126

19.1. Commonly-used symbols 126
19.2. Transistor characteristic curves 129
19.3. Use of the characteristics 134

CHAPTER 20 – TRANSISTOR PARAMETERS 177

20.1. The equivalent circuit of the transistor 177
20.2. The input impedance of the transistor 182
20.3. The output impedance of the transistor 189
20.4. The current gain of the transistor 190
20.5. The internal feedback of the transistor 194

CHAPTER 21 – FORWARD TRANSFER ADMITTANCE OR SLOPE . . 198

21.1. Thermionic valves 198
21.2. Transistors 198
21.3. Derivation of the $-I_C = f(-V_{BE})$ characteristic 201
21.4. Determination of the slope from the $-I_C = f(-V_{BE})$ characteristic . . . 201
21.5. Variation of the slope as a function of the collector current 202

CHAPTER 22 – METHODS OF DRIVING SMALL-SIGNAL TRANSISTORS 213

22.1. Voltage drive 214
22.2. Current drive 222
22.3. The effect of the generator impedance on the internal feedback of the transistor 226

CHAPTER 23 – METHODS OF DRIVING POWER TRANSISTORS 229
23.1. Current drive 229
23.2. Voltage drive 230

CHAPTER 24 – MATCHED DRIVE 233
24.1. Small signals 234
24.2. Large signals 237

CHAPTER 25 – DISTORTION AND POWER AMPLIFICATION 238
25.1. Minimum distortion 238
25.2. Power amplification 241
25.3. Summary 253

CHAPTER 26 – TRANSISTOR BIAS VOLTAGES 258
26.1. Graphical determination of the working point of a transistor 262
26.2. The use of a bias battery 265
26.3. Bias voltage taken from the supply battery 266

CHAPTER 27 – THERMAL STABILITY 269
27.1. Ambient temperature 269
27.2. Collector dissipation 269
27.3. Thermal resistance 270
27.4. Low thermal resistance 277
27.5. Low value of λ 277
27.6. The relationship between the leakage current and the generator impedance 281
27.7. Effects connected with variations of the input characteristic resulting from increased temperatures 282
27.8. Stabilization of the working point of a transistor 289
27.9. Stabilization by means of a resistor between collector and base 293
27.10. Stabilization by a resistor connected in series with the emitter 298
27.11. Base connected to a voltage divider, combined with the use of an emitter resistor 306
27.12. Voltage divider with thermistor 314

CHAPTER 28 – THE LOAD IMPEDANCE 318
28.1. Constant-power hyperbola 318
28.2. Maximum output power of an amplifier 321
28.3. Maximum power gain 330
28.4. The effect of the load impedance on the voltage gain 333
28.5. The effect of the load impedance on the current gain 334
28.6. The effect of the load impedance on the power gain 337

CHAPTER 29 – TRANSISTORS CONNECTED IN CASCADE 341
29.1. Transformer-coupled stages 341
29.2. *RC*-coupled stages 345

CHAPTER 30 – THE USE OF TRANSISTORS AT RADIO FREQUENCIES. 349

CHAPTER 31 – THE TRANSISTOR AS AN ACTIVE ELEMENT 350
31.1. Current-amplification factor 353
31.2. Cut-off frequency 354
31.3. Slope 357
31.4. Determination of the optimum value of the constant collector current . 361

CHAPTER 32 – THE TRANSISTOR AS A PASSIVE ELEMENT 365
32.1. Equivalent circuit 365

CHAPTER 33 – THE INFLUENCE OF THE COLLECTOR CURRENT AND OF THE FREQUENCY ON THE VARIOUS R.F. PARAMETERS OF THE TRANSISTOR 369
33.1. The input circuit 369
33.2. The output circuit 373
33.3. The feedback circuit 381
33.4. Application of the curves 385
33.5. Neutralization 401
33.6. Maximum power gain 402
33.7. The effect of the dependence of the transistor parameters on current, voltage and frequency 405
33.8. Expressions for the circuit elements occurring in the equivalent circuit diagrams. 414

CHAPTER 34 – POSSIBLE CIRCUIT CONFIGURATIONS FOR TRANSISTORS 417

CHAPTER 35 – THE COMMON-EMITTER CONFIGURATION. 418
35.1. Power gain 418
35.2. The influence of the load impedance on the input impedance of the transistor 425
35.3. The influence of the generator impedance on the output impedance of the transistor. 426

CHAPTER 36 – THE COMMON-BASE CONFIGURATION 429
36.1. Power gain 429
36.2. The influence of the load impedance on the input impedance of the transistor 437

36.3. The influence of the generator impedance on the output impedance of the transistor 438

CHAPTER 37 – THE COMMON-COLLECTOR CONFIGURATION 441
37.1. Power gain 441
37.2. The influence of the load impedance on the input impedance of the transistor 449
37.3. The influence of the generator impedance on the output impedance of the transistor 450

CHAPTER 38 – COMPARISON OF THE VARIOUS CONFIGURATIONS 453
38.1. The voltage gain and its dependence on the load impedance 454
38.2. The current gain and its dependence on the load impedance 457
38.3. The power gain and its dependence on the load impedance 458
38.4. The input impedance and its dependence on the load impedance 461
38.5. The output impedance and its dependence on the generator impedance . 462
38.6. Conclusions 465

APPENDIX 466

PART ONE

Physical phenomena in semiconductors

CHAPTER 1

General considerations

1.1 Characteristic properties of semiconductors

Semiconductors are distinguished from other materials (conductors or insulators) by the following properties:

a. Resistivity

The resistivity of a conductor is of the order of 10^{-6} to 10^{-5} Ω.cm, that of an insulator on the other hand is of the order of 10^{6} to 10^{8} Ω.cm, while the resistivity of a semiconductor lies between 10^{-3} and 10^{7} Ω.cm.
The resistivity of a conductor increases very slightly in proportion to the temperature (Fig. 1). On the other hand, the resistivity of a semiconductor decreases exponentially with increasing temperature. (Fig. 2). As a result, in contrast to a conductor, a semiconductor will conduct better when hot than when cold (negative temperature coefficient).

b. Photoconductivity

By photoconductivity we understand the property by which a body offers a smaller or larger resistance to the passage of an electric current, depending on the illumination to which it is exposed. Photoconductivity can be demonstrated with the aid of the circuit given in Fig. 3. Conductors do not show this property at all (Fig. 4); on the other hand, the resistivity of a semiconductor decreases exponentially with the illumination applied to it. (Fig. 5).

c. Rectification

Rectification is that property of a body whereby it offers a low resistance to an electric current in one direction and a high resistance in the other

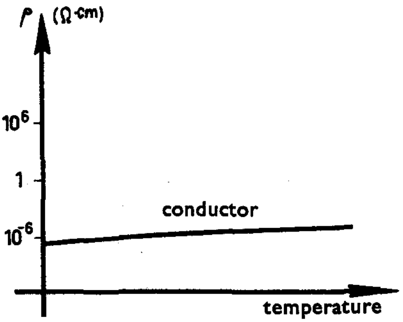


Fig. 1

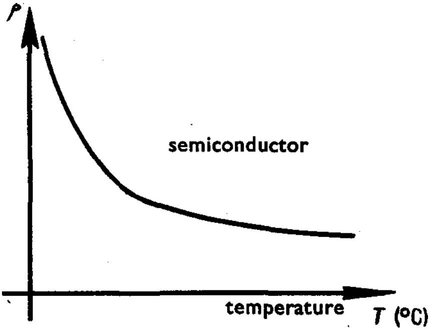


Fig. 2

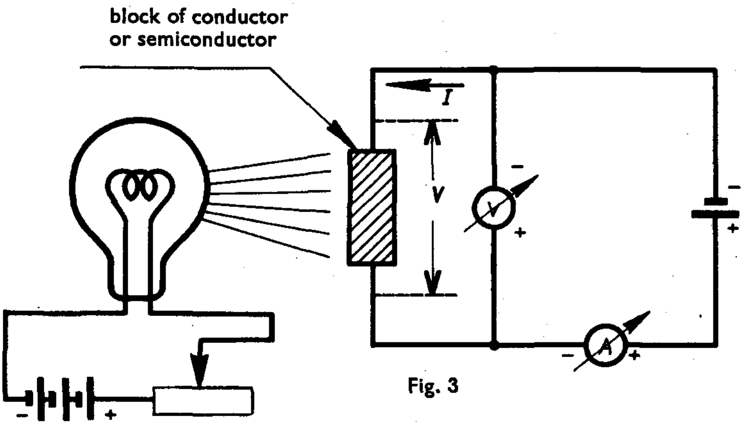

block of conductor
or semiconductor
I
V
V
A
-
+
-
+
-
+
-
+

Fig. 3

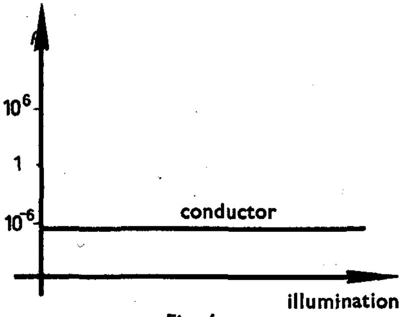


Fig. 4

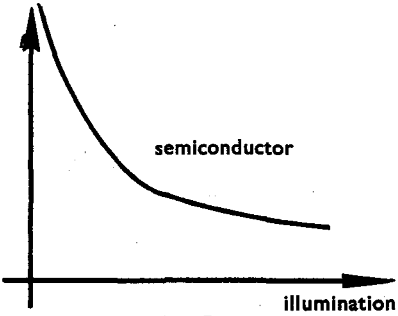


Fig. 5

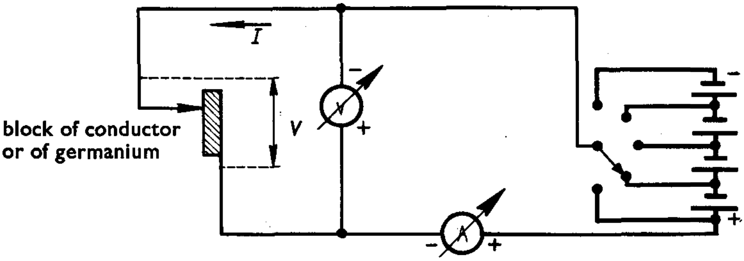

I
block of conductor or of germanium
V
V
A
−
+
−
+
−
+

Fig. 6

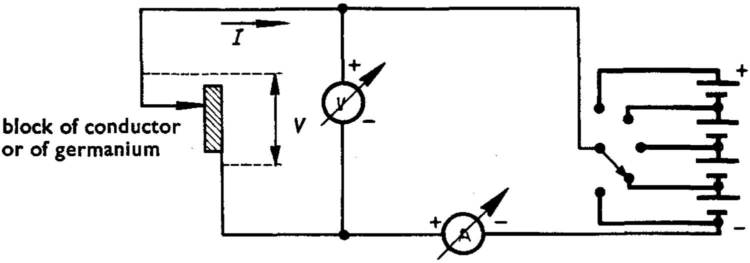

I
block of conductor or of germanium
V
V
A
+
−
+
−
+
−

Fig. 7

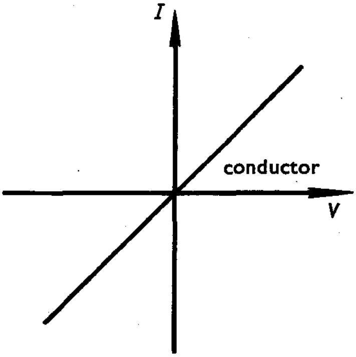


Fig. 8

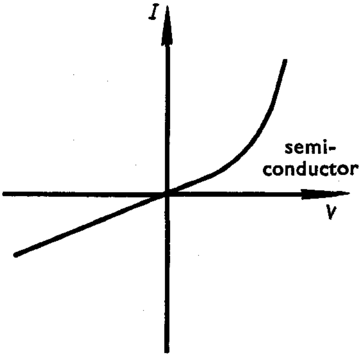


Fig. 9

direction. This property can be demonstrated with the aid of the circuits in Figs. 6 and 7.

A metal point contact on a conductor always results in the conductivity in both directions being equal (Fig. 8), but a metal point contact on a semiconductor will give different conductivities, depending on the direction of the voltage which is applied to the poles. (Fig. 9).

CHAPTER 2

Definition

Semiconductors belong to groups of elements which have a low conductivity, but whose conductivity increases in the presence of certain impurities, and similarly increases with temperature.

2.1. Structure

Let us consider a germanium atom (Ge). This can be regarded as consisting of a nucleus, with electrons describing circular paths around it, distributed in 4 shells, as in Fig. 10: 2 electrons in the first shell, 8 electrons in the second shell, 18 electrons in the third shell, 4 electrons in the fourth shell. The atom formed in this way is electrically neutral, that is, the negative charges to be ascribed to the electrons completely cancel the positive charges on the nucleus.

For the sake of clarity we will represent the germanium atom by a nucleus round which move only the four electrons of the fourth shell (Fig 11). These four valency electrons form the four covalent bonds of the germanium atom. This is also true of some other elements, such as carbon (in the crystalline form occurring in diamonds) and silicon.

2.2. Bonds between two germanium atoms

Let us assume that *A* and *B* are two germanium atoms. Fig. 12 shows that the bond between these two atoms is obtained by the linkage of the two valency electrons a_1 and b_3. Thus we see that each bond takes up one valency electron from each atom. As a germanium atom only has four valency electrons, it can only be linked with four other germanium atoms, and all these will be at the same distance from the first atom and from each other.

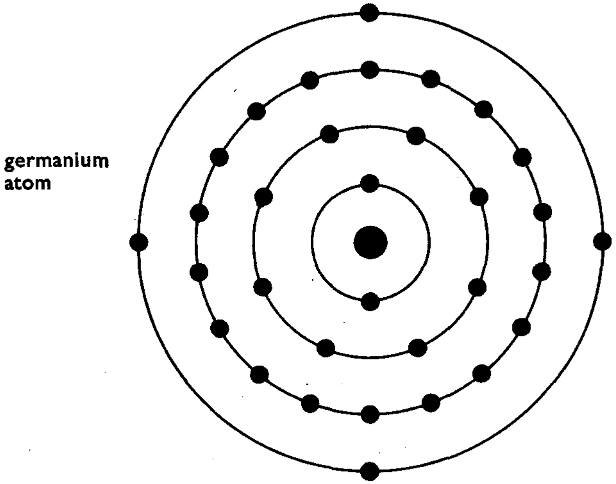

germanium
atom

Fig. 10

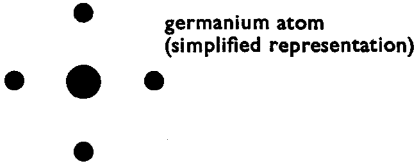

germanium atom
(simplified representation)

Fig. 11

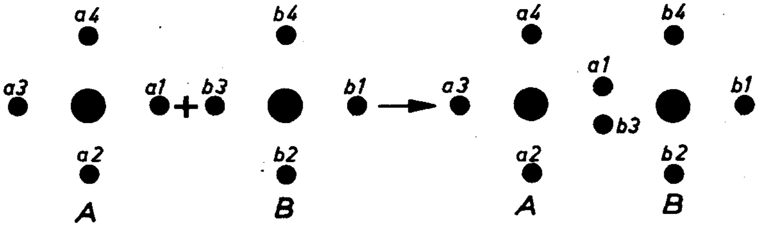

a4
a3
a1
a2
A
b3
b4
B
b1
b2
a4
a3
A
a2
a1
b3
b4
B
b1
b2

Fig. 12

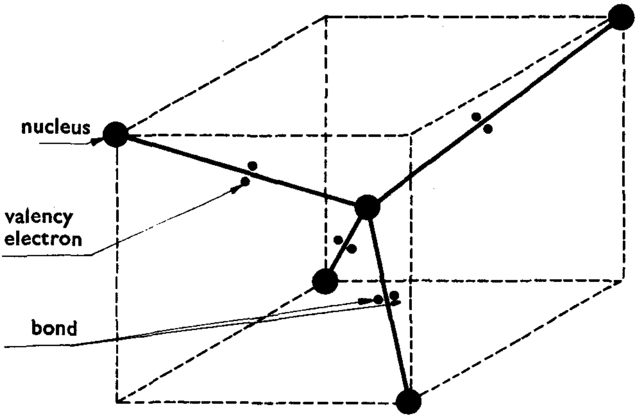

nucleus
valency electron
bond

Fig. 13

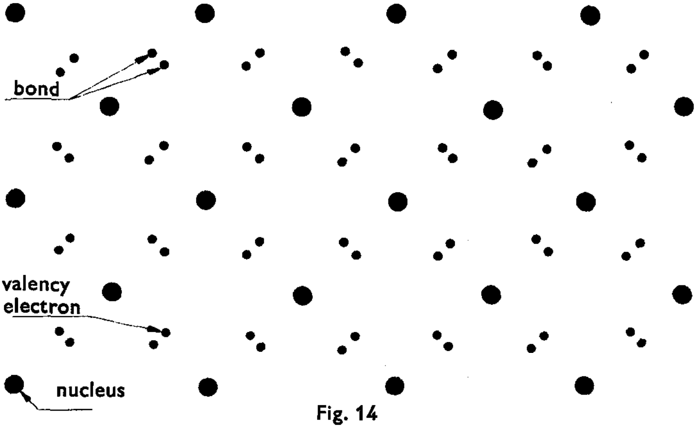

bond
valency electron
nucleus

Fig. 14

2.3. Molecular structure

When all possible bonds, each formed by a pair of electrons, have been completed, the crystal structure is complete. (Fig. 13). For the sake of clarity, we will represent the crystal structure in a simple manner in one plane (Fig. 14). In this figure, the nuclei are represented by the large dots, the valency electrons by the small dots, and the bonds by two linked valency electrons. The figure shows clearly how an atom is linked to each of the neighbouring atoms by one of these valency electrons, so that all the bonds are present, and all the charges on the nucleus are neutralized.

We see that there are no free "charge carriers" (free electrons), so that it is impossible for an electric current to flow. The germanium crystal can thus be regarded as a perfect insulator. The above considerations refer to phenomena which occur at a certain temperature, that is at absolute zero (–273°C); at this temperature there is complete equilibrium.

2.4. Ways in which equilibrium can be disturbed

The equilibrium can be disturbed, either by raising the ambient temperature above –273° C, or by the energy which is radiated by a source of light, or by radiation of a still higher frequency (X-rays, α-, β-, γ-radiation, etc.).

2.5. Increase of temperature

Heat causes molecular vibrations which tend to disturb the arrangement of the atoms. This thermal energy causes the atoms to vibrate, which means that the nuclei move alternately away from each other and towards each other; separation of the atoms can lead to a bond being broken. This results in the liberation of an electron, and the disappearance of the bond. The electron becomes a free electron, while the germanium atom is one negative charge short. (Disappearance of the valency electron). We will represent the absence of the negative charge as the presence of a positive charge, termed a "hole".

This effect is represented in Fig. 15. The free electrons make it possible for an electric current to flow; the conduction of electric current by "holes" is not so easy to represent.

2.6. Conduction by holes

The holes appear to move in the opposite direction to the free electrons, and at about the same velocity.

Let us assume that an external voltage source produces an electric field in the crystals. (Fig. 16). This field forces a valency electron (*b*) which has been liberated by the thermal effect, to "fall" into a "hole" (*A*); the electron recombines with the hole represented by the incomplete bond, and a new hole (*B*) is produced. The same field forces the valency electron (*c*) to fall into the hole (*B*), thus bringing about the recombination of bond (*B*) which has just been broken, while a new hole (*C*) is produced. This process develops further and further (valency electron (*d*) and broken bond (*D*)). The movement of the holes is thus due to the restoration of bonds between atoms.

Fig. 16 shows that the movement of a hole really amounts to the movement of valency electrons in the direction of the electric field. The holes appear to move in the opposite direction to that of the electric field to which they are exposed.

Note

The reader's attention is drawn to the fact that in contrast to the conventional method of regarding electric current in which it is assumed that current flows from the positive pole to the negative pole outside the voltage source, the direction of the current in semiconductors is understood as the direction of the electron current, which flows from the negative pole to the positive pole outside the voltage source.

2.7. Conductivity in a block of germanium

The conductivity of a block of germanium depends on the number of free electrons and the number of holes which are present, and on the mobility of the electrons and holes. There is always the possibility that free electrons will recombine with incomplete bonds, so that the number of free electrons and holes depends on the equilibrium between the rate at which such

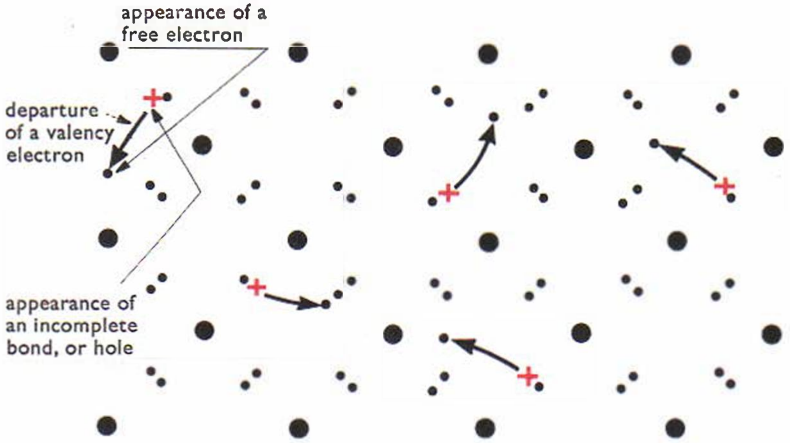

appearance of a free electron
departure of a valency electron
appearance of an incomplete bond, or hole

Fig. 15

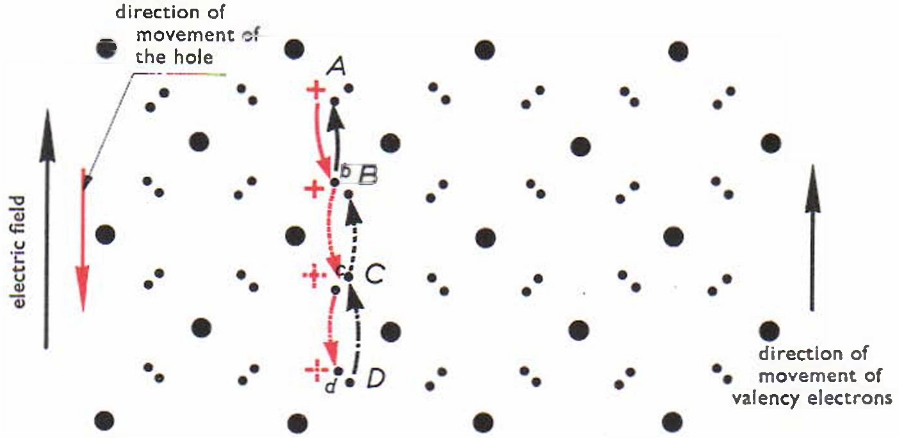

direction of movement of the hole
electric field
A
B
b
C
c
D
d
direction of movement of valency electrons

Fig. 16

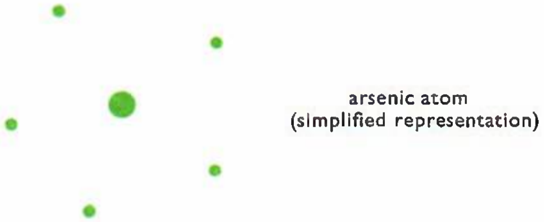

arsenic atom
(simplified representation)

Fig. 17

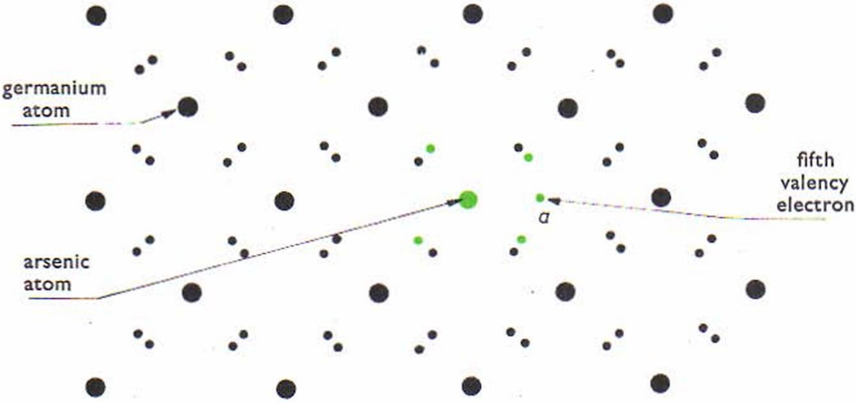

germanium atom
fifth valency electron
a
arsenic atom

Fig. 18

electron-hole pairs are formed, and that at which they recombine. The rate at which free electrons and holes are formed depends on the temperature, and/or the effects of radiant energy, while the rate of recombination is determined amongst other things by the free electron density and the hole density. Equilibrium is attained when the rate of formation of free electrons and holes is equal to the rate of recombination.

At a given temperature the number of free electrons and mobile holes is constant; this number increases with the temperature. As these electrons determine the conductivity of the crystal, this also increases with temperature. As will be explained later, the negative temperature coefficient can give rise to difficulties (instability) at higher temperatures. The conductivity of a block of pure germanium is small. In order to increase it, certain impurities can be introduced into the germanium in order to encourage the production of holes. This does not result in bonds being broken, so that no electrons are liberated in the process.

We will now take a closer look at *N*-type germanium, as it is called, into which an impurity has been introduced which encourages the production of free electrons, and at *P*-type germanium, which is contaminated in such a way as to encourage the formation of holes.

2.8. *N*-type germanium

Let us consider an atom of arsenic. Its geometrical dimensions are identical with those of a germanium atom, so that it can be fitted perfectly into the crystal structure of a block of germanium. As arsenic is pentavalent, we represent it by a nucleus round which five valency electrons move in the fourth shell (Fig. 17). If we introduce an arsenic atom into a block of germanium, (see Fig. 18, in which the arsenic atom is shown in green, and the germanium atoms in black), we see that the impurity atom has one electron too many. Within the crystal structure of the germanium, the arsenic atom can only be linked to the germanium atoms by means of four valency electrons, so that the fifth electron, (*a*) remains linked to its nucleus. However, this bond is very weak. At absolute zero temperature, (–273° C), *N*-type germanium is also a perfect insulator, since it does not contain a single charge-carrier (free electrons or holes). At room temperature, however, the thermal motion is sufficient to remove electron (*a*) from the nucleus, and to turn it into a free electron (Fig. 19).

The conductivity of *N*-type germanium depends on the electron density. Electrons are produced in two ways:

1) By incorporation of impurity atoms into the germanium crystal, whereby each of these atoms liberates one electron.
2) By breaking the bonds between germanium atoms.

The number of free electrons which are formed by incorporation of impurity atoms depends on the extent to which the germanium is "doped" by arsenic; by contrast, the number of free electrons and holes which are produced by the breakage of bonds between germanium atoms, is determined by the temperature.

At room temperature all the impurity atoms have lost their fifth valency electron; it is true that the conductivity of a block of *N*-type germanium increases with temperature, but to a lesser extent than that of pure germanium. In this case, the impurity atoms are termed "donor" atoms or simply "donors". They do in fact give up one electron each to the block of germanium.

Under normal conditions the arsenic atom is electrically neutral, because the negative charge associated with the electrons is then equal to the positive charge on the nucleus. However, if the arsenic atom has lost its fifth valency electron, the state of equilibrium has been disturbed; the arsenic atom has lost part of its negative charge, and has become positively ionized. These donor atoms are firmly held in the crystal lattice and cannot move about. As they have lost their fifth valency electron at room temperature, the donors are electrically positive. The total positive charge on all the donors is equal to the total negative charge on all the electrons which have been liberated, so that the block of *N*-type germanium as a whole is electrically neutral. This type of material is termed *N*-type germanium, because its conductivity must be ascribed to negative charge carriers (free electrons). *N*-type germanium is represented as shown in Fig. 20.

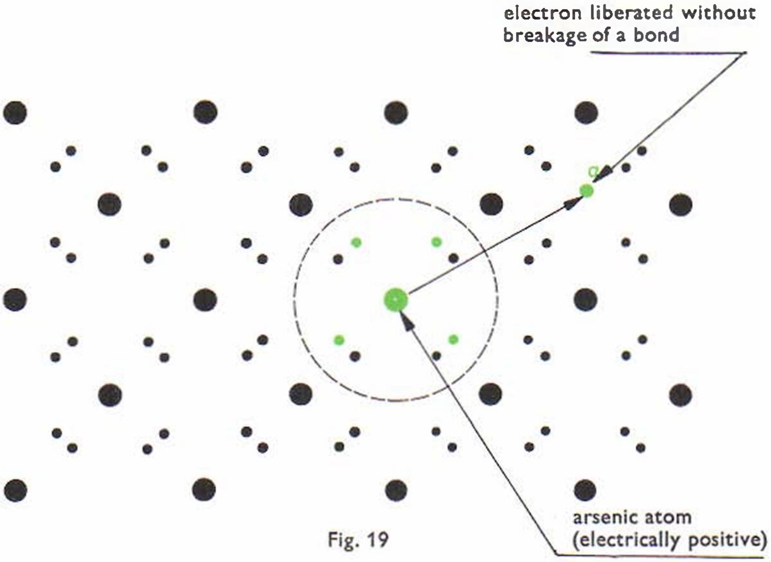

electron liberated without breakage of a bond
a
arsenic atom (electrically positive)

Fig. 19

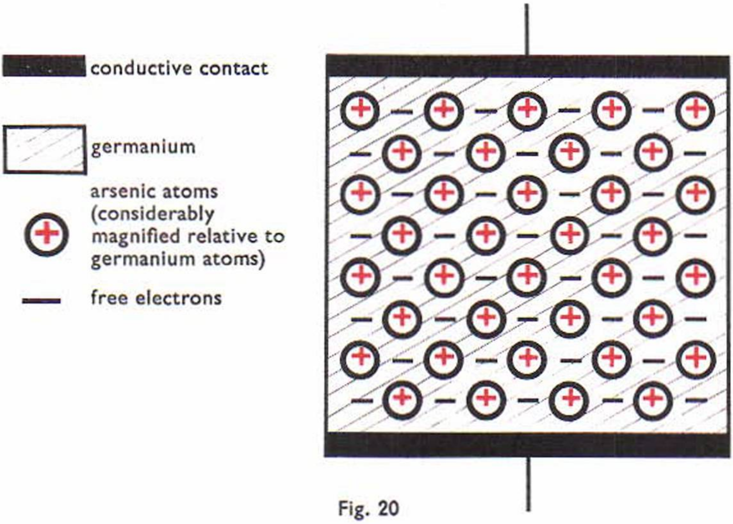
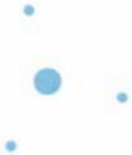

conductive contact
germanium
arsenic atoms (considerably magnified relative to germanium atoms)
free electrons

Fig. 20

Fig. 21


indium atom
(simplified representation)

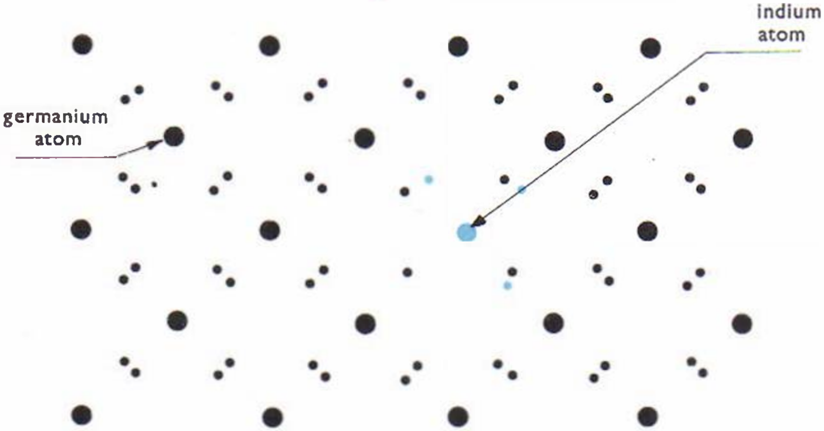

indium
atom
germanium
atom

Fig. 22

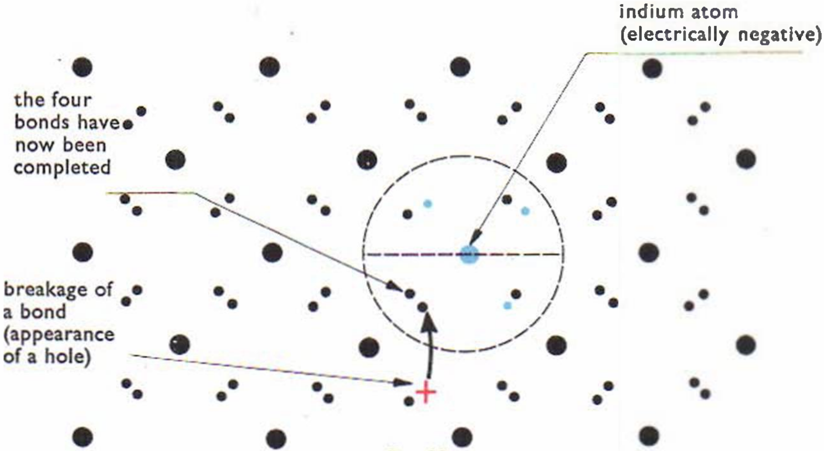

indium atom
(electrically negative)
the four
bonds have
now been
completed
breakage of
a bond
(appearance
of a hole)
+

Fig. 23

2.9. *P*-type germanium

We will now consider an atom of indium. Just as with an arsenic atom, the geometrical dimensions of an indium atom are identical with those of a germanium atom. Consequently it can be completely integrated into the crystal structure of a block of germanium. Since indium is trivalent, we can represent it as a nucleus round which three valency electrons circle in the fourth shell (Fig. 21).

If an indium atom is introduced into a block of germanium, we have the situation which is represented in Fig. 22, in which the indium atom is shown in blue and the germanium atoms in black. The indium atom becomes incorporated in the structure of the crystal, but can only complete three bonds with germanium atoms. This incomplete structure gives rise to a hole. At room temperature, thermal movement liberates some of the valency electrons which provide the bonds between germanium atoms These electrons fall into the holes caused by the presence of indium atoms, with the result that new holes are produced (broken bond) at the place from which the liberated electrons came. This effect is propagated continuously. (Fig. 23).

The conductivity of *P*-type germanium depends on the hole density, which can be produced in the following two ways:

1) by the incorporation of impurity atoms into the germanium crystal, one hole being produced for each impurity atom.
2) By the breaking of bonds between germanium atoms.

The number of mobile holes which are produced by incorporation of impurity atoms depends on the extent to which the germanium is doped by indium, while the number of free electrons and holes resulting from the breakage of bonds between germanium atoms depends on the temperature. At room temperature, all the impurity atoms have gained a fourth electron, and the conductivity of the *P*-type germanium varies with temperature in the same way as that of *N*-type germanium. In this case the impurity atoms are termed "acceptor-atoms" or simply "acceptors". They do in fact accept one electron each. The indium atom is originally electrically neutral, since the negative charge on the electrons is equal to the positive charge on the nucleus.

As soon as a bond is formed between the indium atom and a fourth valency electron from a germanium atom, and the indium atom has accepted a fourth valency electron, the balance of charges is disturbed. The indium atom now possesses an extra negative charge, and is negatively ionized. These acceptor atoms are held fast in the crystal structure and cannot move about. If they have obtained a fourth valency electron at room temperature, they become electrically negative.

The block of *P*-type germanium itself is electrically neutral, because the total negative charge on all the acceptor atoms is equal to the total positive charge on all the mobile holes.

The material is termed *P*-type germanium because its conductivity must be ascribed to what appear to be positive charge carriers (mobile holes). *P*-type germanium is represented as shown in Fig. 24.

2.10. Representation of a block of pure germanium

At room temperature, a block of pure germanium is represented schematically as shown in Fig. 25. Note particularly the presence of a number of free electrons and of an equal number of mobile holes.

2.11. Conduction in a block of *N*-type germanium

We will now investigate the process of conduction in a block of *N*-type germanium with the aid of the circuit of Fig. 26.

a) Switch *S* open

The block of *N*-type germanium is again represented by means of the symbols explained above, that is, the electrically positive arsenic atoms and the free electrons have been exaggerated in size relative to the germanium atoms.

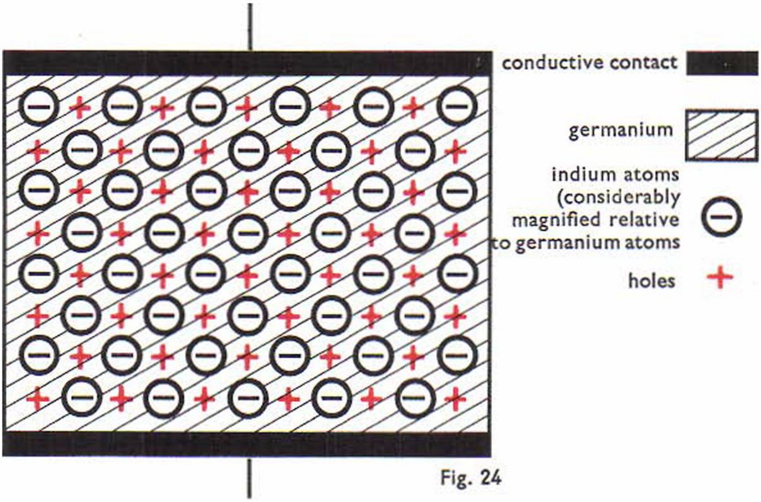

conductive contact
germanium
indium atoms (considerably magnified relative to germanium atoms)
holes

Fig. 24

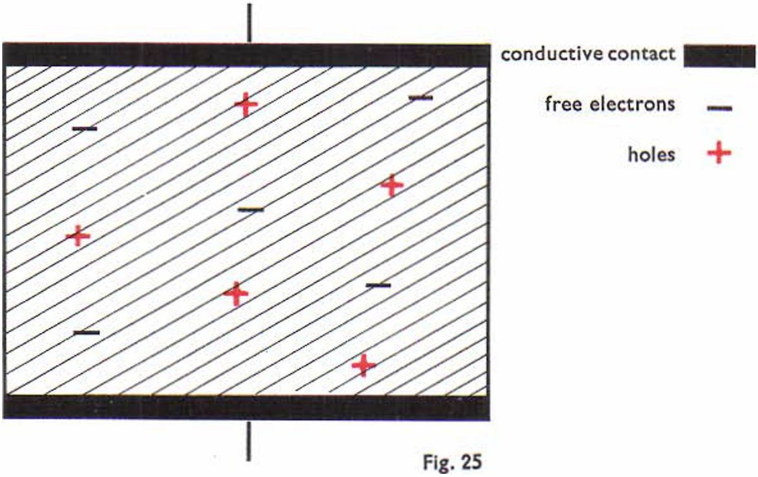

conductive contact
free electrons
holes

Fig. 25

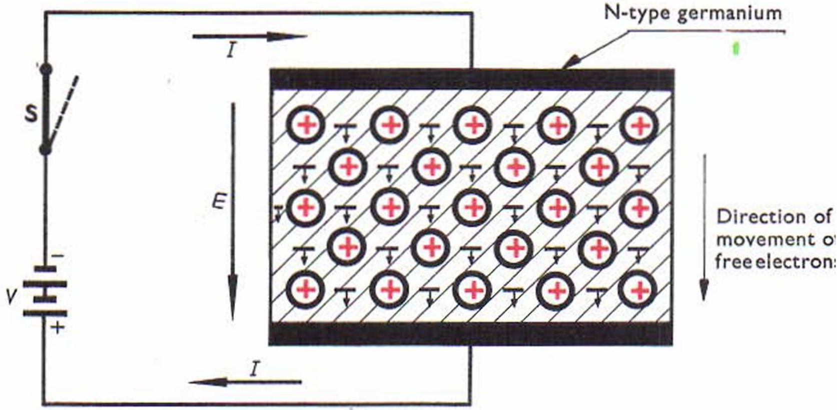

N-type germanium
I
S
E
V
–
+
I
Direction of
movement o
free electron

Fig. 26

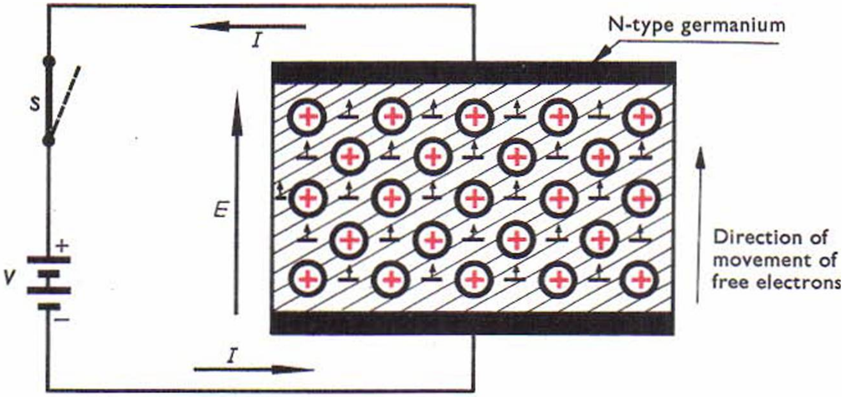

N-type germanium
I
S
E
V
+
–
I
Direction of
movement of
free electrons

Fig. 27

b) Switch *S* closed

The voltage *V* causes an electric field in the block of germanium. If we represent this field as a force which acts from the external negative pole to the positive pole of the battery, and is of amplitude *E*, the free electrons in the block of *N*-type germanium will move in the direction of this force. Current is conducted by the movement of free electrons in the direction of the applied electric field. The conductivity of the material increases with the number of free electrons (percentage of impurity), so that outside the battery the (electron) current flows from negative to positive.

C) Reversal of the polarity of the battery

If the polarity of the battery is reversed as shown in Fig. 27, the electric field is also reversed. The free electrons will now move in the opposite direction, and the electric current outside the battery will again flow from negative to positive at the same voltage. The amplitude of this current will again be equal to that which flowed in the previous case. Consequently, it appears that no rectification whatever occurs in a block of *N*-type germanium.

2.12. Conduction in a block of *P*-type germanium

To investigate the process of conduction in a block of *P*-type germanium, we consider Fig. 28.

a) Switch *S* open

The block of *P*-type germanium is again represented by means of the familiar symbols, that is with disproportionately large negative atoms and mobile holes.

b) Switch *S* closed

The block of *P*-type germanium is now subject to an electric field which we will represent by a force of amplitude *E* acting inside the crystal from negative to positive.

We will now examine the phenomena which occur inside the crystal. We will do this on the basis of Fig. 29, which represents a small piece of *P*-type germanium. The bond of which electron (*b*) forms part is broken, and under the influence of the field the electron fills the hole (*A*). At the same time the liberated electron (*c*) falls into the hole (**B**); conduction within the crystal is evidently due to the movement of the electrons in the direction of the applied electric field (see "Conduction by holes" Page 16). This conduction by valency electrons can be represented as conduction by positive holes, whereby the latter move in the opposite direction to that in which the valency electrons move, that is in the opposite direction to the electric field. (The hole has moved from bond (*A*) to (*C*).) At side I, the electrons which flow into the semiconductor crystal from the connecting lead fill up the existing holes in the crystal. The opposite effect occurs at side II; the holes come in contact with the electrode and are filled up by free electrons, (coming from the electrode itself and from the leads connected to it), which are supplied by the voltage source.

Conduction in a block of *P*-type germanium increases with the number of holes (percentage of impurity); outside the battery the electric current thus flows from the negative pole to the positive pole. To simplify further explanation, we will represent this as a movement of mobile holes from the positive electrode I to the negative electrode II.

e) Reversal of the polarity of the battery

Reversal of the polarity of the battery gives the situation shown in Fig. 30. The electric field is reversed, and the holes move in the opposite direction to that in which they moved in the previous case. The electron current outside the battery again flows from the negative to the positive pole, the magnitude of this current being the same as that which flowed in the previous case at the same voltage.

Consequently, no rectification whatever occurs in a block of *P*-type germanium.

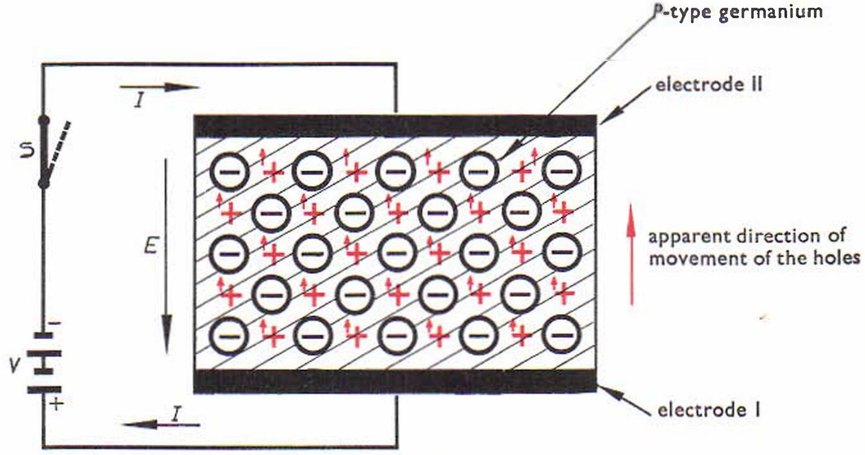

P-type germanium
electrode II
apparent direction of movement of the holes
electrode I
I
I
E
S
V
–
+

Fig. 28

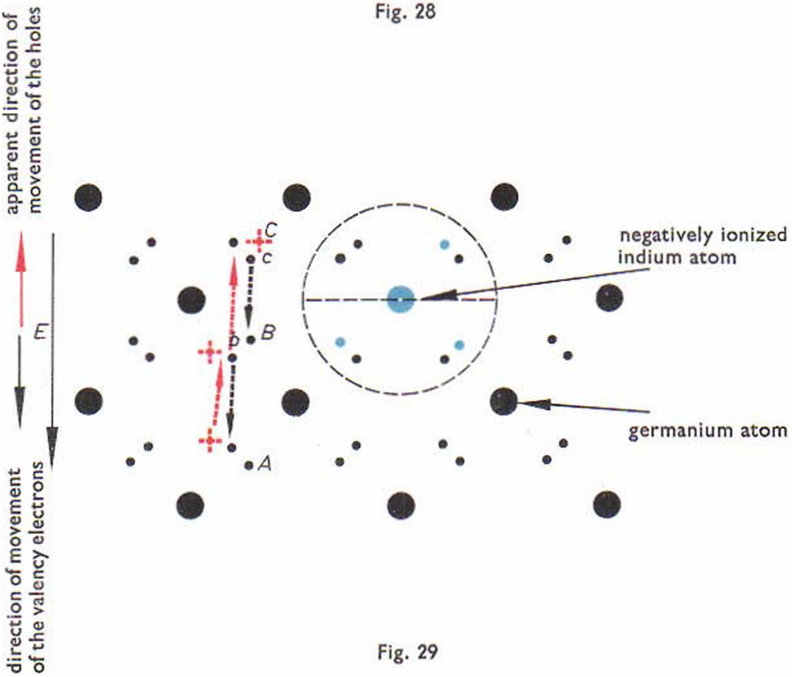

apparent direction of movement of the holes
direction of movement of the valency electrons
E
C
c
B
b
A
negatively ionized indium atom
germanium atom

Fig. 29

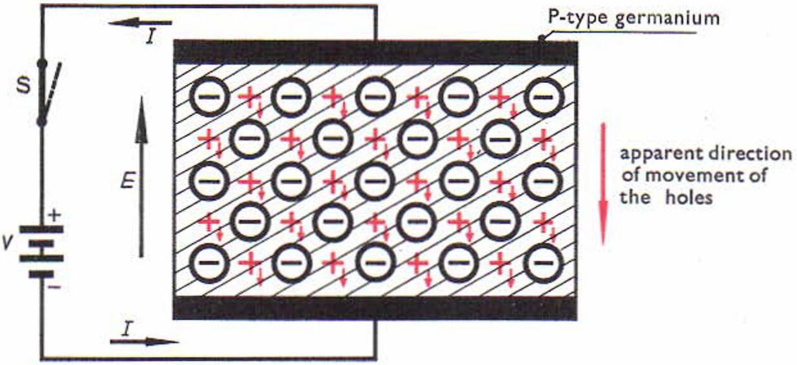

P-type germanium
I
S
E
+
V
−
I
apparent direction of movement of the holes

Fig. 30

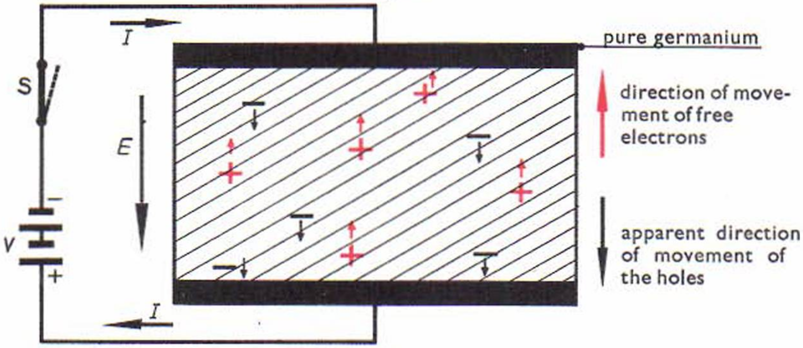

I
pure germanium
S
E
−
V
+
I
direction of movement of free electrons
apparent direction of movement of the holes

Fig. 31

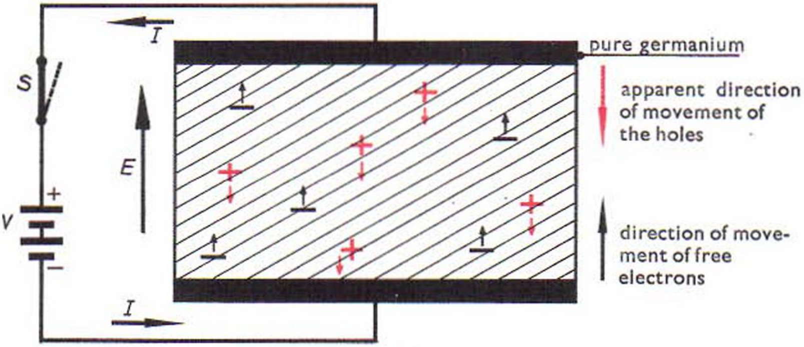

I
pure germanium
S
E
+
V
−
I
apparent direction of movement of the holes
direction of movement of free electrons

Fig. 32

2.13. Conduction in a block of pure germanium

Fig. 31 represents the circuit for this case.

a) Switch *S* open

At a certain temperature, the block of germanium contains free electrons and mobile holes. Relatively speaking, there are far fewer charge carriers – free electrons and mobile holes – than there are in *N*-type germanium or *P*-type germanium.

b) Switch *S* closed

The battery *V* produces an electric field in the block of germanium. The free electrons will move in the direction of this field, and the holes in the opposite direction; consequently a current will flow round the circuit. With the same voltage, this current will be much smaller than that in the previous cases, because the conductivity of the block of germanium depends of course on the number of charge carriers present. Conduction is due, to an equal extent, to both holes and free electrons.

If we reverse the polarity of the battery, a current of the same size will flow through the circuit, but in the opposite direction. (Fig 32). Consequently, just as in a block of *N*-type germanium or *P*-type germanium, no rectification effect whatever occurs in a block of pure germanium. It is evident, then – as far as conduction is concerned – that neither direction is preferred.

2.14. The *PN*-junction

a) Definition

A *PN*-junction in a single crystal is a thin region in which the conduction changes from type *P* to type *N*.

In principle, a junction of this type can be obtained as follows:

- by allowing impurities to diffuse in a semiconductor.
- by alloying with *P*- or *N*-type elements.

If a germanium crystal is partly *P*-type and partly *N*-type, there must of necessity be a region between the two different types of material, in which conduction by means of holes changes to conduction by means of free electrons.

b) Absence of an external voltage

Let us examine a single crystal which is composed of two blocks of germanium of different types (see Fig. 33). In the *N* region the number of free electrons will be greater than the number of holes, and in the *P* region the number of holes will be greater than the number of free electrons. The free electrons will diffuse from the *N* region to the *P* region, while the holes will diffuse from the *P* region to the *N* region.

In this way, the free electrons coming from the *N* region arrive in the region where the number of holes is relatively large, so that they will recombine with the holes very rapidly. On the other hand, the holes coming from the *P* region will arrive in a region where the number of free electrons is relatively large, and they will very quickly become filled by the electrons in the *N* region. As the donor and acceptor atoms occupy fixed positions, the result will be a movement of negative charges from the *N* region to the *P* region, and of positive charges in the opposite direction. (Fig. 34).

The permanently negative ions (electrically negative indium atoms), which form an integral part of the crystal lattice, constitute a fixed negative space charge. The permanently positive ions (electrically positive arsenic atoms) form a fixed positive space charge.

c) Diffusion voltage or potential barrier

This double space-charge layer causes an electric field which is due to internal factors; this field is directed from the negative space charge (*P*-type germanium) to the positive space charge (*N*-type germanium), and is of magnitude *e* (Fig. 35a).

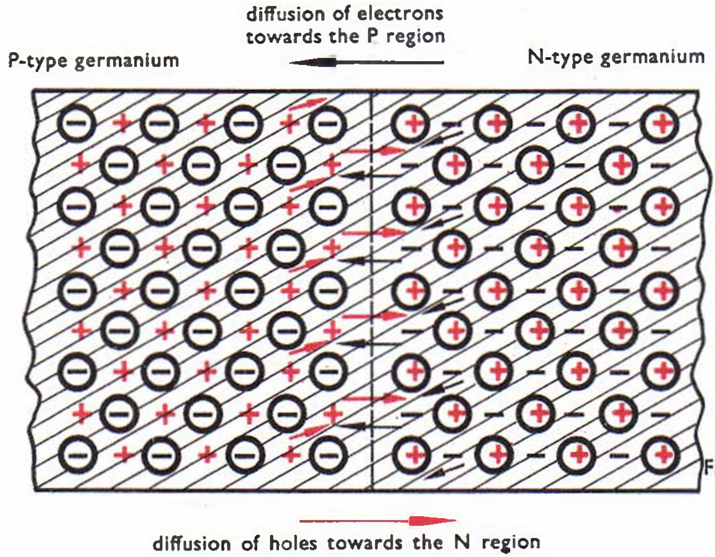

diffusion of electrons towards the P region
P-type germanium
N-type germanium
diffusion of holes towards the N region

Fig. 33

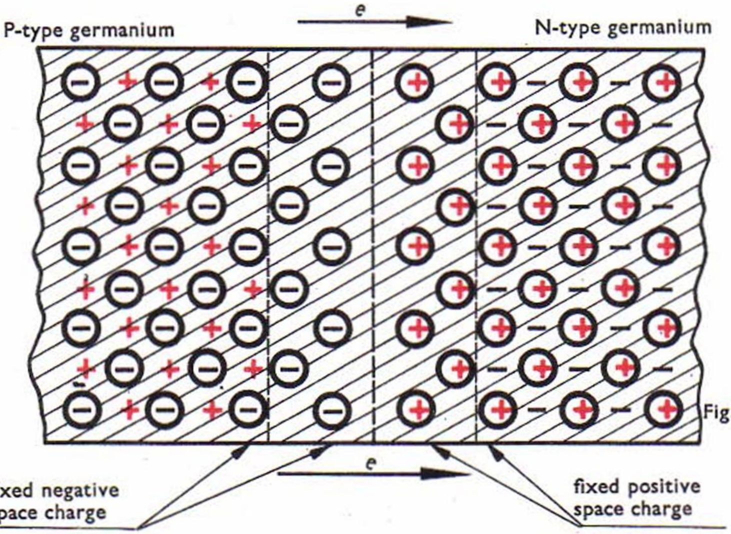

P-type germanium
e
N-type germanium
e
fixed negative space charge
fixed positive space charge

Fig. 34

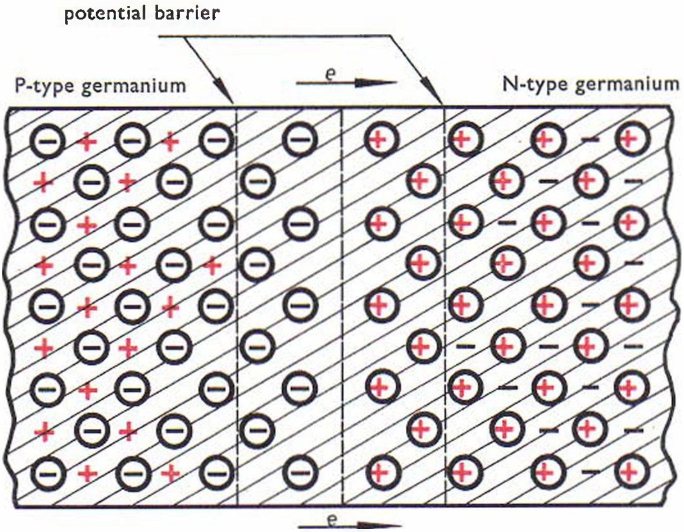

potential barrier
P-type germanium
e
N-type germanium
e

Fig. 35a

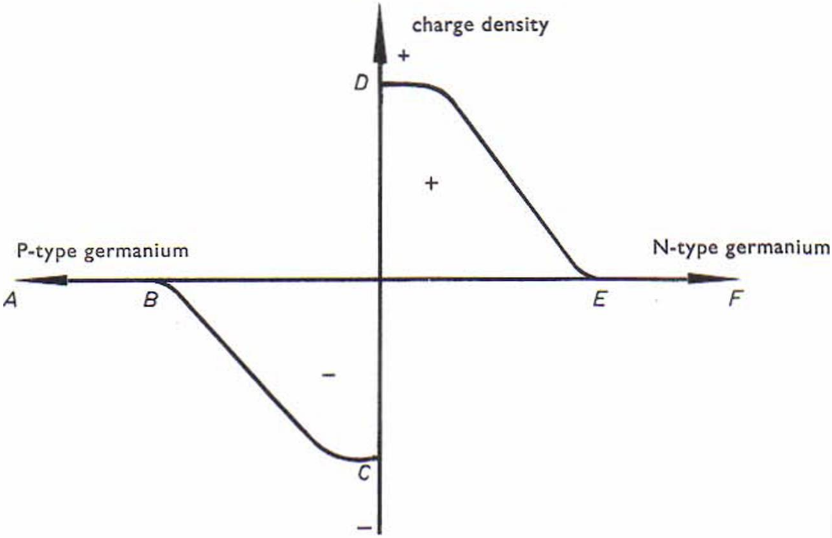


Fig. 35b

The electric field *e* hinders the diffusion of holes into the *N* region and of free electrons into the *P* region; it forms a barrier which drives the free electrons back into the *P* material and the electrons into the *N* material. This barrier is termed a "potential barrier".

The mobile charge carriers in the immediate neighbourhood of the junction move away from it. This narrow region, which is about one micron thick, can be regarded as a transition zone which is limited by a potential barrier.

d) Charge density

Fig. 35b represents the charge density in a germanium crystal containing a *PN*-junction, as a function of the distance from the junction. This curve can be explained as follows.

That part of the *P*-type germanium region which is furthest away from the junction does not receive a single free electron from the *N* region, so that the equilibrium of the electric charges is maintained; this region is always electrically neutral (*AB*, Fig. 35b).

In the neighbourhood of the junction, on the *P*-type germanium side, the electrons from the *N* region have filled up a number of holes; this disturbs the electrical equilibrium and the extent of the disturbance increases with the number of holes in the *P* region which are filled up by free electrons from the *N* region. The charge density is a negative maximum in that region of the transition zone (on the *P*-type germanium side) where there is no longer a single mobile hole (point *C*, Fig. 35b). The same reasoning is applicable to the *N* region; the charge density is a positive maximum in that region of the transition zone (on the *N*-type germanium side) where there is no longer a single free electron (point *D*, Fig. 35b). Between points *C* and *D* there is a sudden transition from the maximum negative charge density to the maximum positive charge density.

The portion *EF* of the curve corresponds to that region of the *N*-type germanium where no holes have been able to diffuse from the *P* region.

2.15. The junction capacitance

Only the fixed electric charges remain in the transition zone. The ionized atoms are held in fixed positions which they cannot leave, because they form part of the crystal lattice; in the *P*-type germanium they are acceptors and in the *N*-type germanium, donors. There are no longer any mobile charge carriers present in this zone, so that it can be regarded as a perfect insulator.

This effect enables us to ascribe a characteristic capacitance to the junction, the thickness of the dielectric being determined by the width of the junction. The value of this capacitance is given by the expression:

$$C = KS/4\pi d$$

in which K represents the permittivity, S the surface area and d the thickness of the junction.

2.16. The effect of an external electric field on the junction

a. In the forward direction

We shall base our considerations on the following circuits.

Switch S open (*Fig.* 36*a*)

The internal electric field *e*, which is directed from the *P*-type germanium towards the *N*-type germanium, hinders free electrons from moving into the *P* region and mobile holes from moving into the *N* region. As the switch is open, no electric current can flow through the junction.

Switch S closed (*Fig.* 36*b*)

The applied voltage *V* produces an electric field which, outside the battery, is directed from the negative pole to the positive pole. This electric field *E* opposes the electric field *e*; *E* encourages the diffusion of charge carriers so that the electrons move from the *N* region to the *P* region, while holes move from the *P* region to the *N* region. In this way a current *I* will flow in the circuit, which means that the junction is conductive.

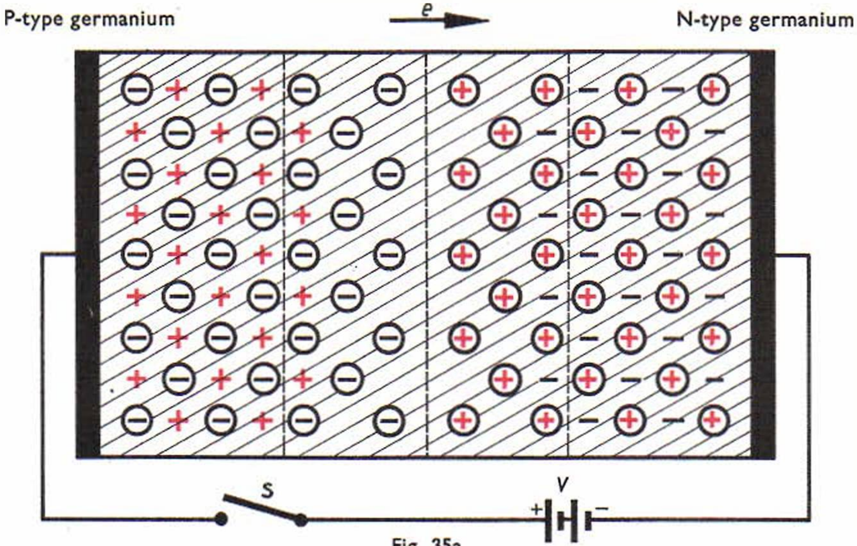

P-type germanium
e
N-type germanium
s
V
+
−

Fig. 35a

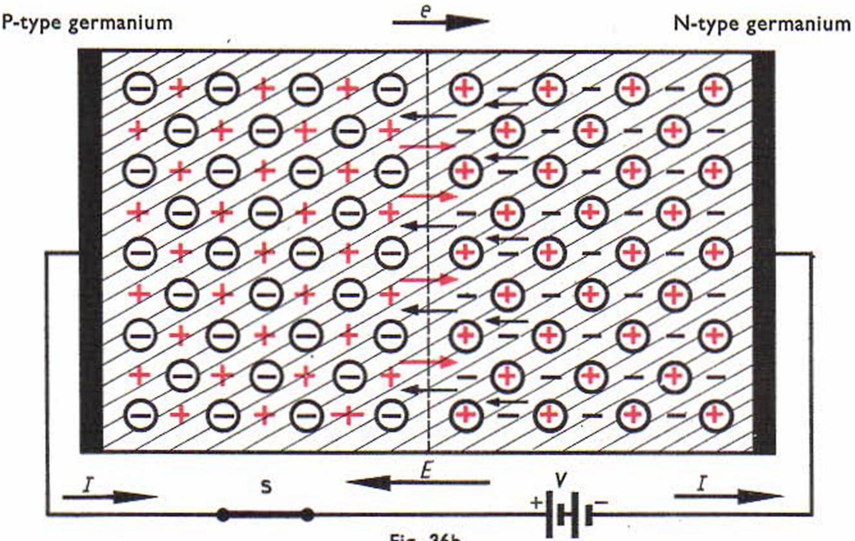

P-type germanium
e
N-type germanium
I
s
E
V
+
−
I

Fig. 36b

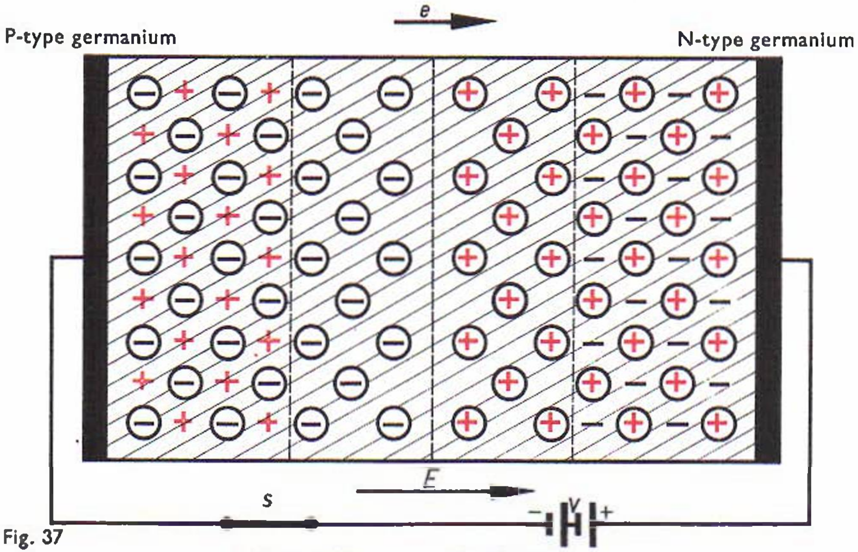

P-type germanium
e
N-type germanium
E
S
V

Fig. 37

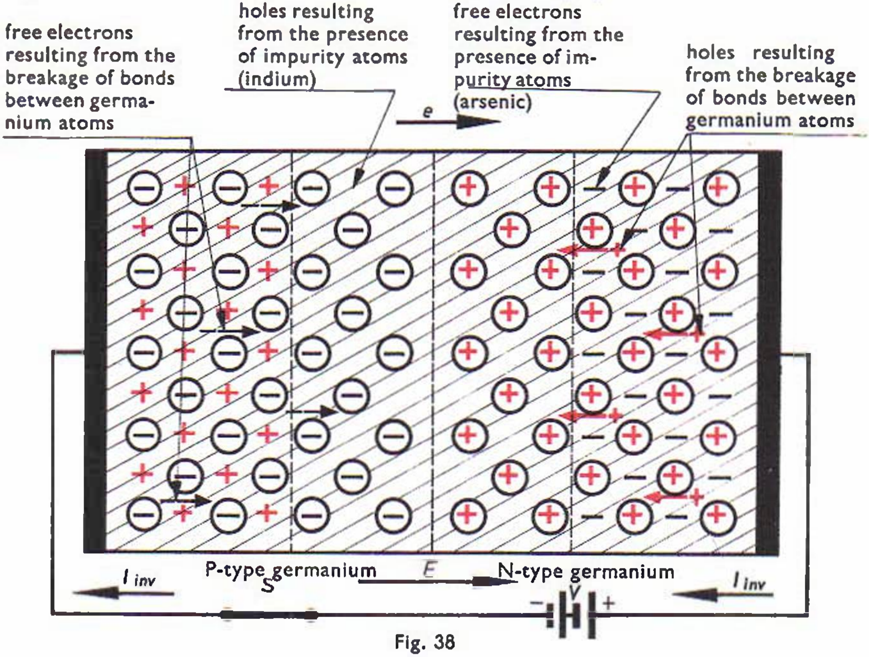

free electrons resulting from the breakage of bonds between germanium atoms
holes resulting from the presence of impurity atoms (indium)
free electrons resulting from the presence of impurity atoms (arsenic)
holes resulting from the breakage of bonds between germanium atoms
e
I inv
P-type germanium
S
E
N-type germanium
V
I inv

Fig. 38

b. In the reverse direction

We will now consider the case when the battery is reversed (Fig. 37). The electric field *E* now acts in the same direction as the internal electric field *e*, and will thus reinforce the effect of the latter. The free electrons from the *N* region cannot diffuse into the *P* region, just as the mobile holes from the *P* region cannot diffuse into the *N* region, which means that the junction is not conductive. In practice, however a small current will still flow through the circuit. It is very important to investigate the cause of this current, which flows when the junction is polarised in the reverse direction so that it should block the flow of current.

In the *N*-type germanium, we have only shown the free electrons which are due to the presence of impurities. Conduction in this type of germanium depends on the following factors:

1) The number of free electrons derived from impurity atoms, which has a very great effect.
2) The breaking of bonds between germanium atoms, which has a smaller effect at room temperature.

As already explained, the breaking of these bonds results in the appearance of free electrons and mobile holes. Consequently a block of *N*-type germanium always contains a small number of mobile holes, which increases with the ambient temperature. In the *P*-type germanium we have only shown the mobile holes which are due to the impurity atoms. Just as in the previous case, however, a small fraction of the conduction is due to the breaking of bonds between germanium atoms, which leads to the production of a few free electrons in the *P* region. Fig. 38 shows that the external electric field *E* permits the few free electrons to move from the *P* region into the *N* region, while the few mobile holes can move from the *N* region into the *P* region, which explains the small current flowing through the circuit. The numbers of free electrons in the *P* region and mobile holes in the *N* region will consequently increase with increasing temperature. This current thus increases with the temperature.

PART TWO

Diodes

CHAPTER 3

Junction diodes

3.1. Principal characteristics

The semiconductor diode is electrically asymmetrical, as is shown by the fact that the current which flows through the diode in one direction is much less than the current which flows through it in the opposite direction. In addition, the characteristic has a sharply curved region, which enables the diode to fulfill the detection function of a thermionic diode.

a) The forward characteristic

The *PN*-junction diode is connected in the forward direction when the *P* region is connected to the positive pole of the battery and the *N* region is connected to the negative pole (Fig. 39). The resulting voltage V_J across the junction is equal to the difference between the battery voltage V_B and the potential drop V_b across the junction, or in other words:

$$V_J = V_B - V_b$$

The junction is polarised in the forward direction, so that a low voltage causes a large current to flow. As the resistance in the forward direction is extremely low, a resistor must be included in the circuit in order to prevent the diode from being damaged. The current-voltage characteristic is an exponential curve (see Fig. 40).

b) The reverse characteristic

In this case, the *P* region is connected to the negative pole of the battery and the *N* region to the positive pole (Fig. 41). The voltage V_J across the junction is equal to the sum of the battery voltage V_B and the potential drop V_b across the junction, or in other words:

$$V_J = V_B + V_b$$

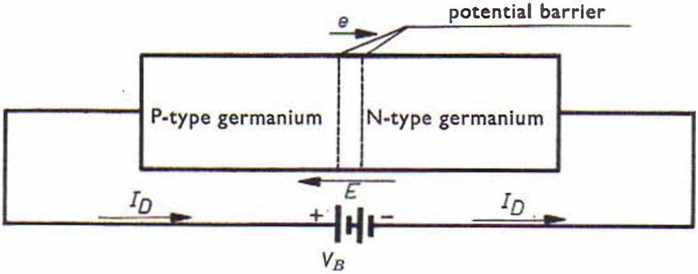

potential barrier
e
P-type germanium
N-type germanium
E
I_D
+
−
I_D
V_B

Fig. 39

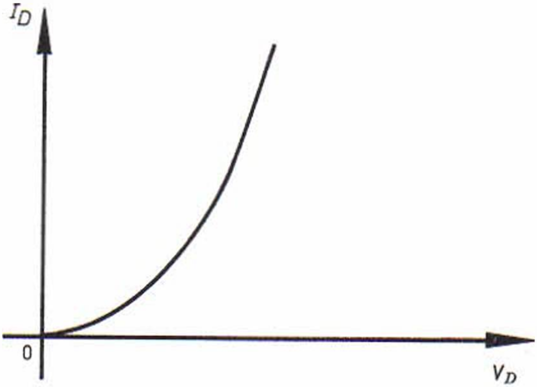


Fig. 40

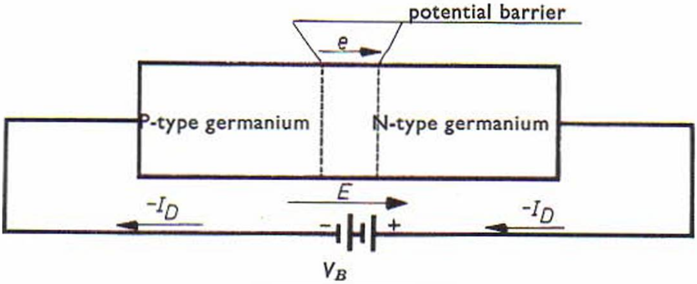

potential barrier
e
P-type germanium
N-type germanium
E
$-I_D$
−
+
$-I_D$
V_B

Fig. 41

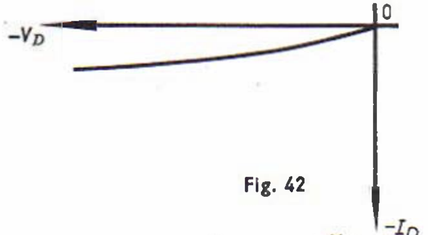


Fig. 42

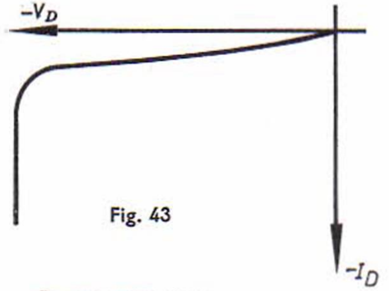


Fig. 43

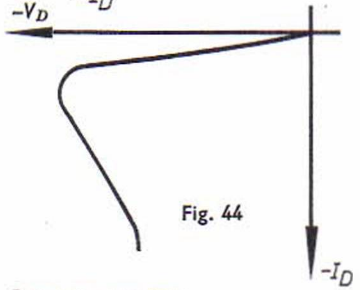


Fig. 44

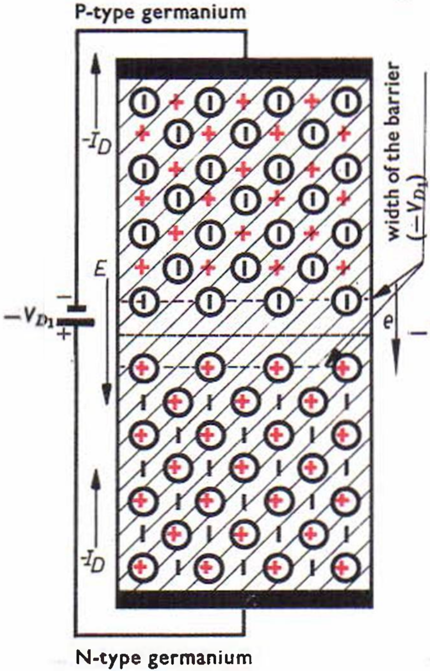

P-type germanium
$-I_D$
E
$-V_{D_1}$
$-I_D$
width of the barrier
$(-V_{D_1})$
e
N-type germanium

Fig. 45

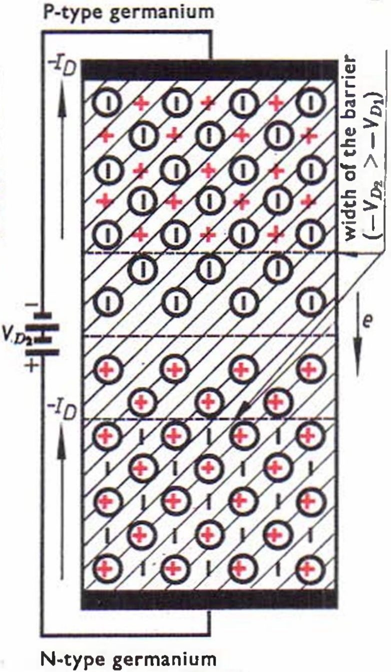

P-type germanium
$-I_D$
$-V_{D_2}$
$-I_D$
width of the barrier
$(-V_{D_2} > -V_{D_1})$
e
N-type germanium

Fig. 46

The current in the external circuit now flows in the opposite direction to its direction in the previous case; this current is very small and increases little with the applied voltage. This reverse current is represented by the symbol $-I_D$ and approaches a saturation value (Fig. 42). The voltage whose polarity is such that the junction blocks the current is termed the "reverse voltage" $-V_D$. At a certain value of the reverse voltage $-V_D$ the curve $-I_D = f(-V_D)$, which represents the current as a function of this voltage, can take two different forms:

1) A straight line, corresponding to a practically constant voltage, the "breakdown voltage" (Fig 43).

2) A curve whose slope becomes negative, after reaching a maximum value of the reverse voltage. (Fig. 44).

c) Variations of the junction capacitance

Previously we have shown that the capacitance of the junction depends on its width (see page 36). If the junction is polarised in the reverse direction, its width increases with the applied voltage (Figs. 45 and 46). This means that its capacitance decreases. In practice, this capacitance may vary between a few picofarads and some tens of picofarads.

CHAPTER 4

Point-contact diodes

PN-junctions may sometimes be the result of non-uniform distribution of impurities during the solidification of a crystal. They can be introduced as required, in either sense, by adding the required type of impurity (indium or arsenic) to the germanium during crystallization.

In semiconductor diodes for certain applications, the *PN*-junction is obtained by lowering a suitable sharp-pointed metal wire onto the crystal, and passing a current through it for a short time (forming the contact). Impurities from the metal wire diffuse from the surface of the crystal into the interior, and in so doing form a *PN* transition zone round the point of contact. This "forming" of the contact in point-contact diodes (Fig. 48) is extremely important; the diode effect does not occur until the contact has been formed. Here too, there is a certain junction capacitance present in the *PN* transition zone. In point-contact diodes, the capacitance of the contacts is relatively small (of the order of 1 picofarad). This is the reason why such diodes offer certain advantages for use at high frequencies. On the other hand they are unsuitable for high powers, so that junction diodes are preferred for such applications.

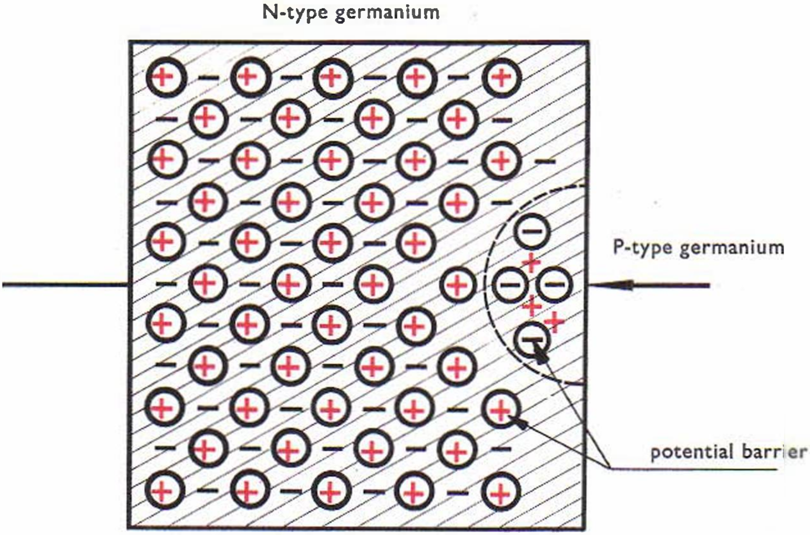

N-type germanium
P-type germanium
potential barrier

Fig. 47

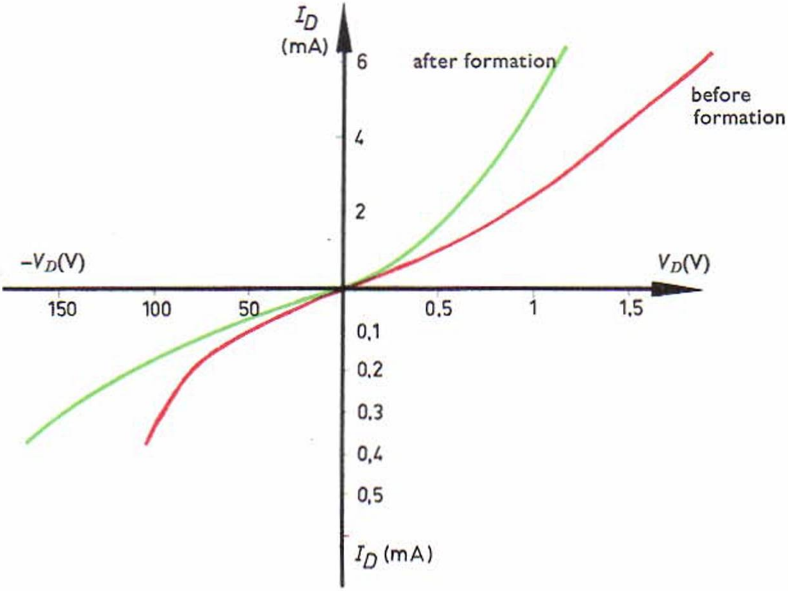

I_D
(mA)
6
4
2
after formation
before formation
$-V_D$(V)
V_D(V)
150
100
50
0,5
1
1,5
0,1
0,2
0,3
0,4
0,5
I_D (mA)

Fig. 48

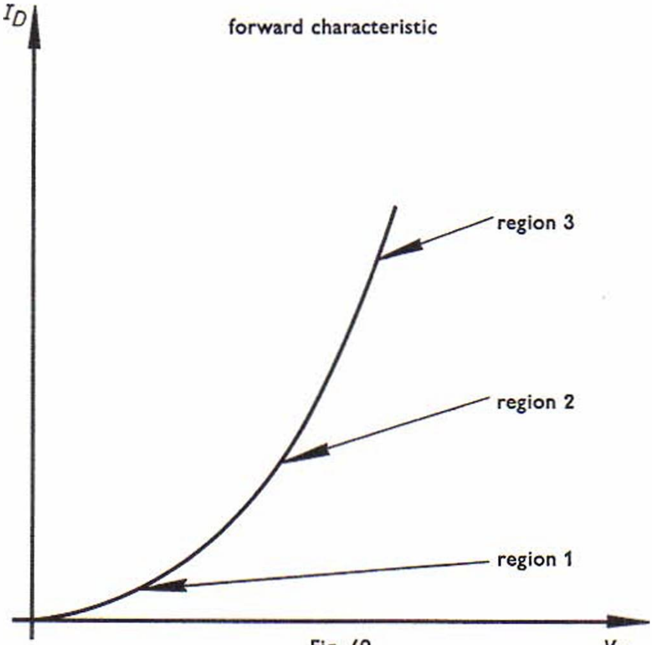


Fig. 49

CHAPTER 5

Analysis of diode characteristics

5.1. Diode biased in the forward direction

In order to bias a diode in the forward direction, we connect the positive pole of the battery to the P region, and the negative pole to the N region. For ease of discussion, we can divide the diode characteristic into three regions (Fig. 49).

a) Region 1

This is the region of very small currents and low voltages (I_D of the order of some tens of microamperes and V_D of the order of some millivolts). The curve is approximately parabolic, except near the origin where it is practically straight.

b) Region 2

In this region the curve is no longer parabolic, but neither is it linear. This region is particularly important in the use of the diode as a detector.

c) Region 3

This is the region of large currents; here, the curve is practically linear. Fig. 49 shows that small variations of voltage in this region result in large variations of current.

d) Resistance

It can be of importance to determine the resistance of the diode in the forward direction, and from this to deduce the variation of this resistance as a function of the voltage. Let us look at the curve given in Fig. 50. Consider the value of the voltage V_D which is represented by point A' on the abscissa. If we draw a perpendicular to the axis at this point, this perpendicular determines point A on the characteristic. The projection of A on the I_D axis gives point B'. We now draw the tangent to the curve at point A, and extend the line $B'A$. The angle between the tangent and the horizontal line through point A, is given by the tangent of this angle:

$$\tan A = \frac{CB}{AB} = \frac{C'B'}{A'B''} = \frac{\Delta I_D}{\Delta V_D} = \frac{1}{R_{D(A)}}.$$

We see that the tangent of this angle is equal to the reciprocal of the resistance of the diode in the forward direction (for a given voltage).

If we increase the value of V_D so that it is represented by the point E' on the abscissa, this point corresponds to point E on the characteristic and point F' on the ordinate. The angle between the tangent to the curve at point E and the horizontal line is now given by the expression:

$$\tan E = \frac{GF}{EF} = \frac{G'F'}{E'F''} = \frac{\Delta I_D}{\Delta V_D} = \frac{1}{R_{D(E)}}.$$

Angle E is greater than angle A, and tan E is thus greater than tan A, from which it follows that $1/R_{D(E)}$ is greater than $1/R_{D(A)}$, and consequently $R_{D(E)}$ is smaller than $R_{D(A)}$. This means that the forward resistance of a diode decreases when the voltage increases; in region 1, this resistance can have a high value, and in region 3 it can become very low.

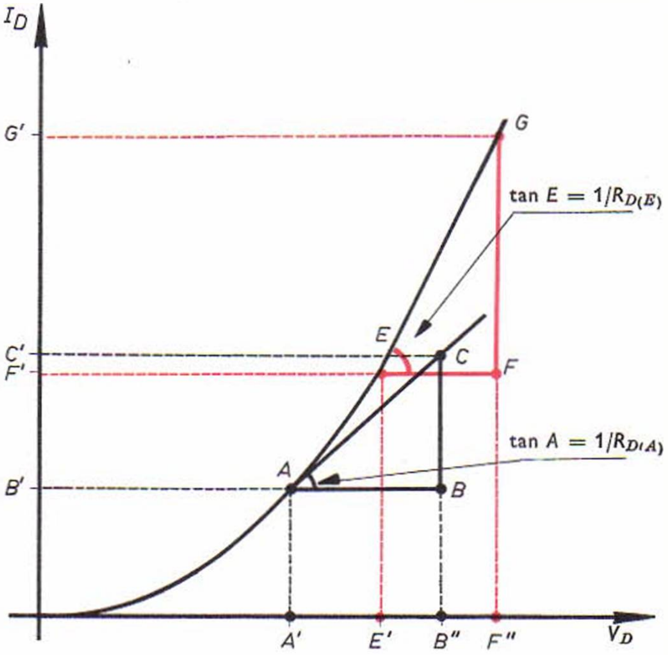

I_D
G'
G
tan E = 1/$R_{D(E)}$
E
C'
F'
C
F
tan A = 1/$R_{D(A)}$
A
B'
B
A'
E'
B''
F''
V_D

Fig. 50

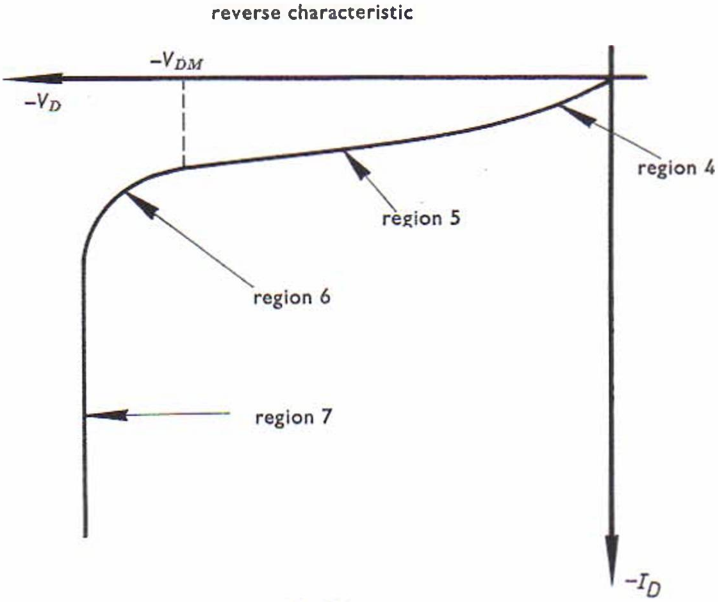

reverse characteristic
$-V_{DM}$
$-V_D$
region 4
region 5
region 6
region 7
$-I_D$

Fig. 51

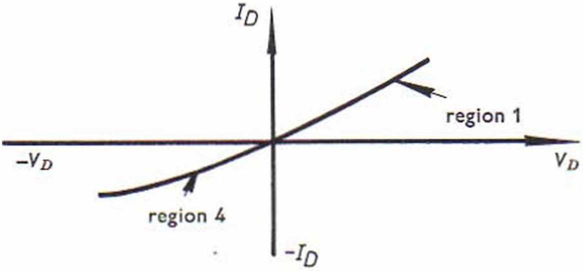

I_D
region 1
$-V_D$
V_D
region 4
$-I_D$

Fig. 52

5.2. Diode biased in the reverse direction

If the positive pole of the battery is connected to the *N*-type germanium, and the negative pole to the *P*-type germanium, the diode is biased in the reverse direction. We can divide the reverse characteristic into four regions.

a) Region 4

This is the region in which both current and reverse voltage have very low values. Once again the curve in this region is practically parabolic, except near the origin, where it is again almost straight. Since region 1 of the forward characteristic is of the same form as region 4 of the reverse characteristic the junction behaves symmetrically in both these regions (Fig. 52). From this, we may conclude that no rectification takes place at very low voltages.

b) Region 5

Here the reverse current approaches a constant value (saturation current), but never reaches it. The resistance in the reverse direction is high, but never becomes infinite.

c) Region 6

In this region the reverse current increases appreciably for a small increase in the reverse voltage. In silicon diodes, this region is extremely narrow, but it is much wider in germanium diodes. The maximum permissible peak value of the reverse voltage $-V_{DM}$ is always lower than the value at which the characteristic begins to bend steeply.

d) Region 7

In this region the "turnover" region breakdown of the junction occurs, and the reverse current rises to a considerable value.

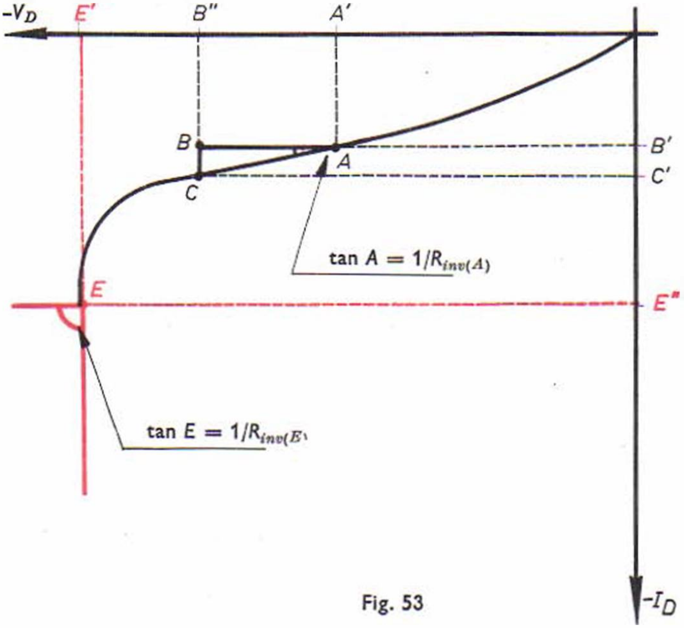
e) The reverse resistance

We wil now determine the reverse resistance of a diode, and investigate how it varies as a function of voltage variations. We will base our considerations on Fig. 53. Consider the value of the reverse voltage $-V_D$ which is represented by point A' on the abscissa. If we draw the perpendicular at this point, it determines point A on the characteristic. The projection of this point on the ordinate gives us point B'. We now draw the tangent to the curve at point A, and extend the line $B'A$. The angle between the tangent to the curve and the horizontal line through point A is given by its tangent:

$$\tan A = \frac{CB}{AB} = \frac{C'B'}{A'B''} = \frac{\triangle I_{D\mathrm{inv.}}}{\triangle V_{D\mathrm{inv.}}} = \frac{1}{R_{\mathrm{inv}(A)}}.$$

We see that the tangent of this angle is equal to the reciprocal of the reverse resistance of the diode (for a given reverse voltage).

If we increase the voltage $-V_D$, so that we are working in region 7 of the characteristic (e.g. point E' on the abscissa) this corresponds to point E on the characteristic and point E'' on the ordinate. We will now determine the angle between the tangent to the curve at point E and the horizontal line through this point. This angle is evidentally almost 90°, so that the tangent of angle E is almost infinite, thus very much greater than the tangent of angle A. From this it follows that $1/R_{\mathrm{inv}\ (E)}$ is very much greater than $1/R_{\mathrm{inv}\ (A)}$, or in other words $R_{\mathrm{inv}\ (A)}$ is very much greater than $R_{\mathrm{inv}\ (E)}$. The reverse resistance of a diode thus increases as the angle between the characteristic and the horizontal decreases. The region in which the characteristic bends downwards can be described as that in which the reverse resistance changes from a very high value to a very low value.



-V_D
E'
B''
A'
B
B'
A
C
C'
tan A = 1/$R_{inv(A)}$
E
E''
tan E = 1/$R_{inv(E)}$
-I_D

Fig. 53

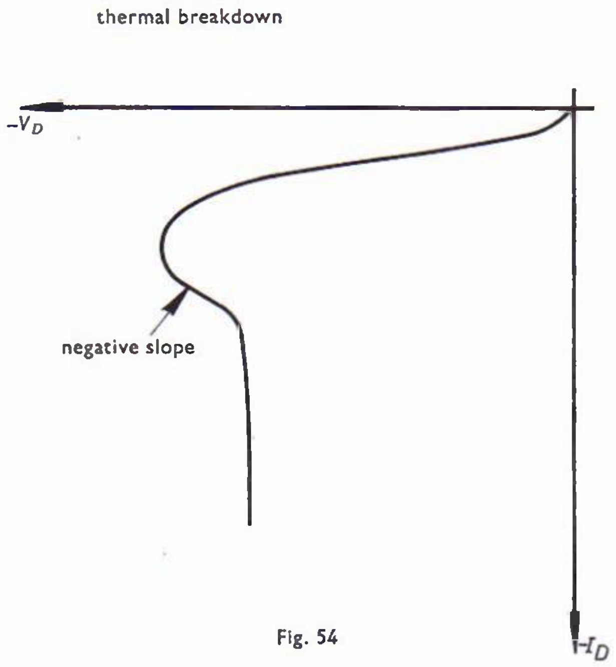


Fig. 54

CHAPTER 6

The breakdown of a junction

The reverse characteristic has a region in which a diode breaks down, that is, a region in which the reverse current becomes very large (region 7, Fig. 51). The breakdown of a diode junction can be ascribed to various effects, i.e.:

1) The thermal effect.
2) The avalanche effect.
3) The zener effect.

6.1. Thermal breakdown

The power dissipated in the junction is equal to the product of the voltage across the junction, and the current which flows through it:

$$P_J = U_{\text{inv}} \,.\, I_{\text{inv}}.$$

The power dissipated in the junction thus increases with the reverse current. We have already explained the origin of the reverse current (page 39), and we found that it increases with temperature. The temperature which the junction reaches depends on:

1) the power which is dissipated in it
and 2) the method by which the diode is cooled.

We must distinguish two cases:

a) the heat dissipated and the cooling are the same, so that the reverse current is stable.
b) The thermal equilibrium is disturbed. The reverse current through the junction increases rapidly with the temperature, and the dissipated heat exceeds the heat that is removed by cooling. This effect is of a cumulative character (Fig. 54), and the characteristic has a negative slope here.

6.2. Avalanche breakdown

The electric charges that flow through the junction receive an amount of energy which is proportional to the voltage across the junction. Every increase in the reverse voltage $-V_D$ corresponds to an increase in the energy which is stored up in the moving charges. As soon as this energy exceeds a certain threshold, electrons are torn loose from germanium atoms as a result of collisions. This effect may also become cumulative, the liberated electrons liberating other charges in their turn. This leads to an avalanche effect, as represented in Fig. 55. In this case, the reverse current is limited only by the components forming the circuit of which the diode forms a part.

6.3. Zener breakdown

The electric field in the junction also increases with the reverse voltage. At a high value of this field breakdown occurs, an effect which is also familiar in heavy-current engineering, and valency electrons are torn loose. The reverse current then increases rapidly, and is no longer limited by the diode itself. In certain cases, the breakdown can be represented by a straight line, which corresponds to a practically constant reverse voltage (Fig. 56). This breakdown voltage can be used as a reference voltage. The "knee" in the characteristic of germanium diodes (Fig. 57) is much more gradual than it is in silicon diodes (Fig. 58). This is the reason why zener diodes, in which this effect is turned to good advantage, are made of silicon.

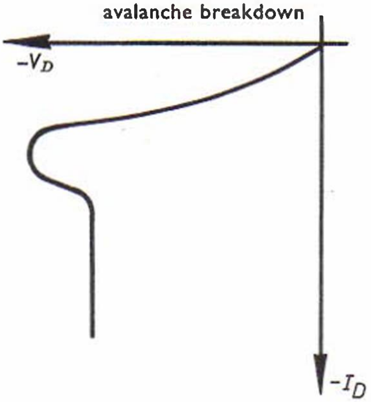

avalanche breakdown

Fig. 55

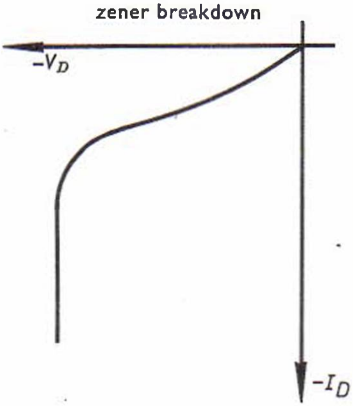

zener breakdown

Fig. 56

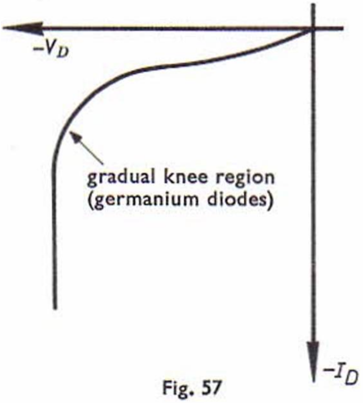

$-V_D$
gradual knee region
(germanium diodes)
$-I_D$

Fig. 57

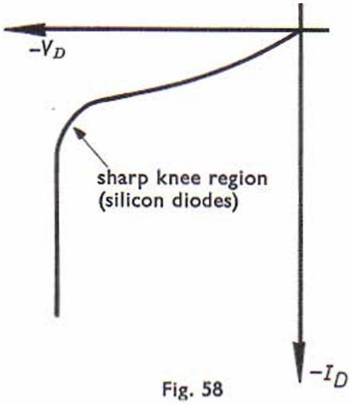
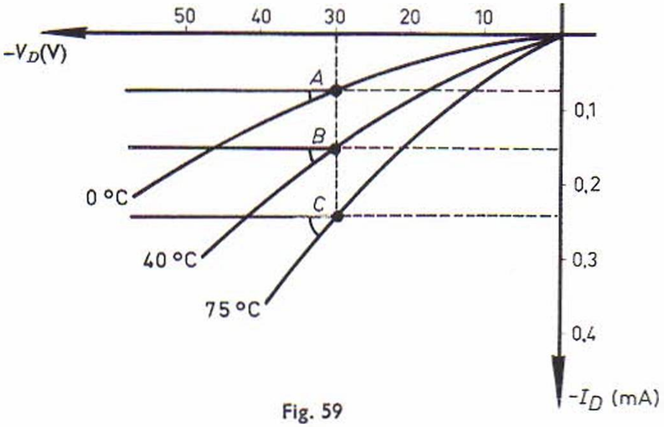
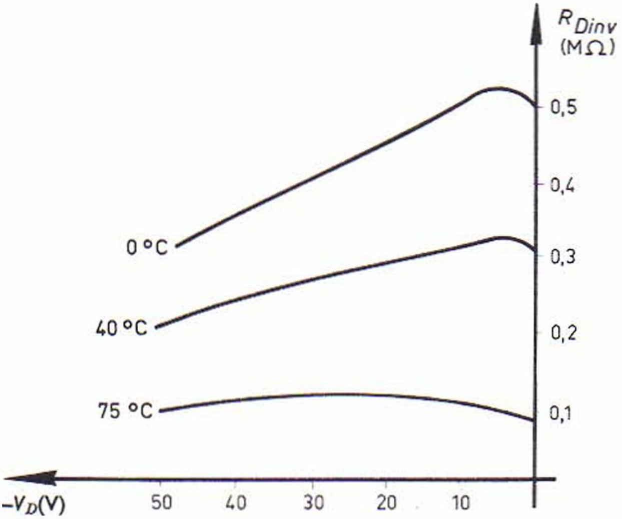

$-V_D$
sharp knee region
(silicon diodes)
$-I_D$

Fig. 58



50
40
30
20
10
$-V_D$(V)
A
B
C
0 °C
40 °C
75 °C
0,1
0,2
0,3
0,4
$-I_D$ (mA)

Fig. 59



R_{Dinv} (MΩ)
0,5
0,4
0,3
0,2
0,1
0 °C
40 °C
75 °C
$-V_D$(V)
50
40
30
20
10

Fig. 60

CHAPTER 7

The influence of temperature

The temperature has a greater influence on the reverse characteristic of a germanium diode than on its forward characteristic.

7.1. The influence of temperature on the reverse characteristic

a) Reverse current

The reverse current increases with the temperature. For a germanium diode, the value of this current doubles if the temperature rises by 10 deg. C; for a silicon diode, it doubles in value for a temperature rise of only 7 deg. C. However, the reverse current of a silicon diode is only one hundredth to one thousandth of that of a germanium diode of the same dimensions. This is the reason why silicon diodes are more suitable for use at higher temperatures. The maximum temperature which a germanium diode can stand is about 75° C, and the maximum temperature for a silicon diode is about 150° C. Fig. 59 represents the change in the reverse characteristic of a germanium diode, if the temperature increases from 0° C to 75° C.

b) Resistance

The curves in Fig. 59 show that for a given reverse voltage the current increases with the temperature. Let us assume that the reverse voltage is 30 V; this corresponds to point *A* on the characteristic for 0° C, to point *B* on the characteristic for 40° C, and to point *C* on the characteristic for 75° C. Angle *C* is larger than angle *B* which in its turn is larger than angle *A*, or in other words $\tan C > \tan B > \tan A$. Since the tangent is equal to the reciprocal of the reverse resistance, we have

$$R_{\mathrm{inv}(C)} < R_{\mathrm{inv}(B)} < R_{\mathrm{inv}(A)}$$

The reverse resistance thus decreases with increasing temperature (Fig. 60).

c) Breakdown voltage

The voltage at which breakdown occurs in a diode as a result of an excessive increase in the crystal temperature, decreases with increasing temperature. (Fig. 61). In a germanium diode the reduction of the breakdown voltage is of less importance than the large increase in the reverse current with increasing temperature. This is because it is impossible to make any use of the breakdown voltage, as the knee of the characteristic bends very slowly, and consequently it is necessary to remain a good way below the breakdown voltage in order to ensure reliable operation of the diode.

7.2. The influence of temperature on the forward characteristic·

The form of the forward characteristic is influenced only slightly by the temperature. This characteristic shifts in a direction parallel to the abscissa, and becomes slightly distorted in doing so (Fig. 63). At a given voltage V_D the current increases slightly with the temperature (point *A* on the characteristic for 0° C and point *B* on that for 40° C.). The forward resistance thus decreases with an increase in temperature. Relatively speaking, the influence of temperature on the forward current and resistance is much less than it is on the reverse current and resistance.

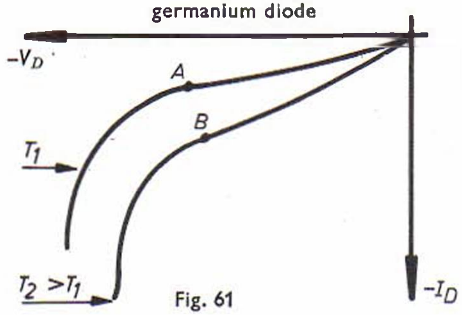


Fig. 61

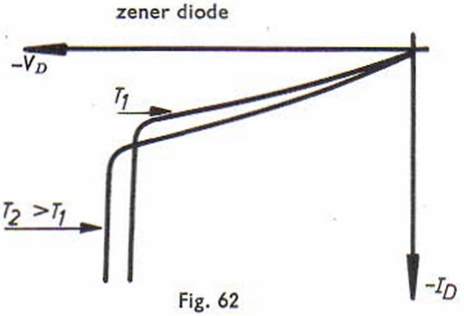


Fig. 62

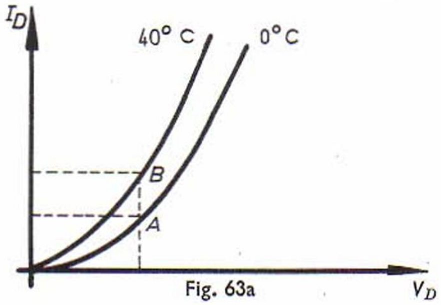


Fig. 63a

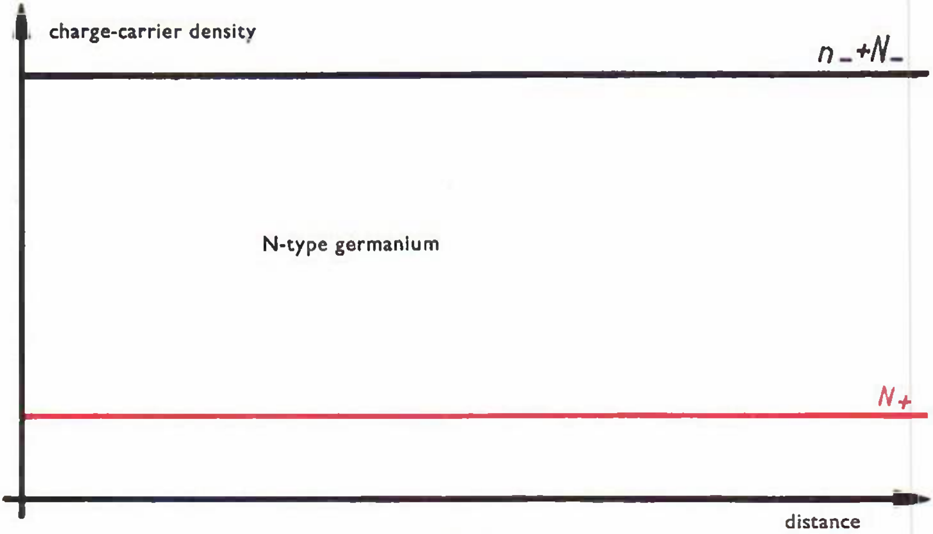

charge-carrier density
$n_- + N_-$
N-type germanium
N_+
distance

Fig. 63b

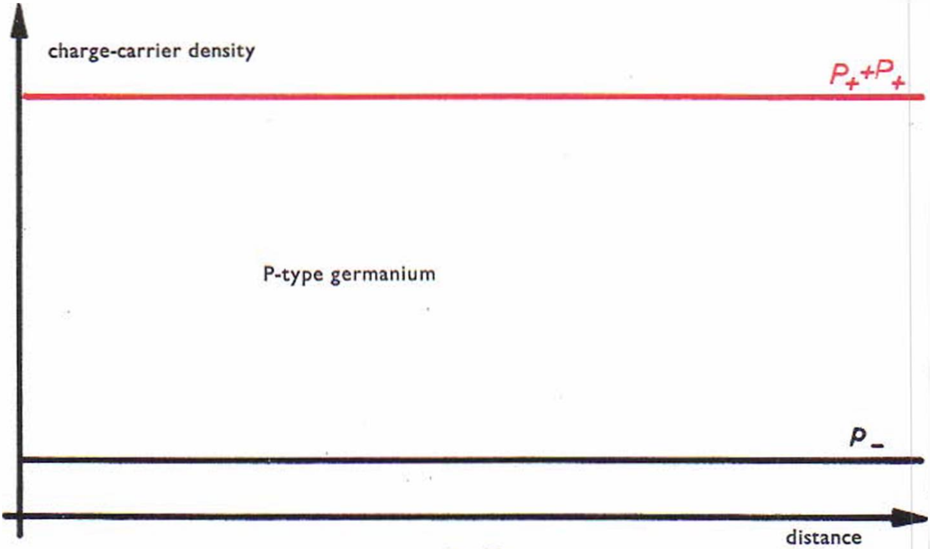

charge-carrier density
$P_+ + P_+$
P-type germanium
P_-
distance

Fig. 63c

7.3. Explanation of the difference in the influence of temperature on the forward and reverse characteristic of a diode

A certain number of impurity atoms are introduced into the pure germanium; arsenic atoms in order to obtain N-type germanium, or indium atoms in order to obtain P-type germanium. Suppose that there are N arsenic atoms present per unit volume of the N-type germanium. At room temperature each of these atoms will have given up an electron so that N represents the density of the free electrons resulting from the presence of the arsenic atoms. At the same time a number of bonds from the germanium atoms will have been broken, giving rise to extra free electrons n_- and mobile holes n_+.

The number of charge-carriers per unit volume of the block of N-type germanium under consideration is thus made up as follows (Fig. 63b):

- free electrons: $N_- + n_-$
- mobile holes : n_+

It must be remembered that N is very much greater than n. Suppose that the number of indium atoms per unit volume of the P-type germanium is P. At room temperature each of these atoms will have given rise to a mobile hole. $P+$ thus represents the density of mobile holes which are due to the presence of the indium atoms. At the same time a certain number of bonds between germanium atoms will have been broken. The number of mobile holes and free electrons per unit volume, which are due to the breakage of these bonds, will be represented by p_+ and p_- respectively.

Consequently the density of charge-carriers in the block of P-type germanium under consideration, is made up as follows (Fig. 63c):

- mobile holes : $P_+ + p_+$
- free electrons: p_-

In the P-type germanium the number of mobile holes P_+ caused by the presence of the indium, is very much greater than the number of mobile holes p_+ resulting from the breakage of bonds between atoms at a given temperature.

7.4. Diode biased in the forward direction

The field E (Fig. 63d) produced by the battery opposes the internal field e, and thus increases the tendency towards diffusion. The free electrons from the N region diffuse into the P region, and the mobile holes from the P region diffuse into the N region. The junction conducts, and so a current will flow through the circuit. This current is a function of the number of charge-carriers which can move as a result of the application of field E. The number of charge-carriers which pass through the junction will depend on the density of free electrons in the N region and on the density of the mobile holes in the P region. Consequently, this current is a function of $N_- + n_- + P_+ + p_+$, that is of $N_- + P_+ + n_- + p_+$.

With an increase in temperature, the number of broken bonds between germanium atoms will increase. Let us represent the new densities of the electrons and holes in the N and P regions by n'_- and p'_+ respectively. At room temperature, all the impurity atoms will have given rise to either a free electron (N-type germanium), or a hole (P-type germanium) so that N_- and P_+ will remain unchanged.

Let us assume that $n'_- = 2n_-$ and that $p'_+ = 2p_+$.

Consequently, the new value of the current will be a function of:

$N_- + P_+ + n'_- + p'_+$ that is of $N_- + P_+ + 2(n_- + p_+)$.

Since N_- and P_+ are very much greater than n_- and p_+ respectively, doubling the latter will only result in a slight decrease in the voltage drop in the forward direction.

7.5. Diode biased in the reverse direction

In this case, the field E reinforces the internal field e (Fig. 63e). Only the holes which are in the minority in the N region can diffuse into the P region, while only the free electrons which are in the minority in the P region can diffuse into the N region. The current is thus a function of $n_+ + p_-$. An increase in the temperature, equal to the increase in the previous example, will result in an increase in the density of the mobile holes in the N region (n'_+) and of the free electrons in the P region (p'_-). The new value of the current is now a function of $n'_+ + p'_-$, or in other words of $2(n_+ + p_-)$. The increase in the temperature thus results in double the number of charge-carriers passing through the junction. For a larger increase in temperature, this number will increase even more sharply, which clearly demonstrates the importance of the influence of temperature changes on the value of the reverse current.

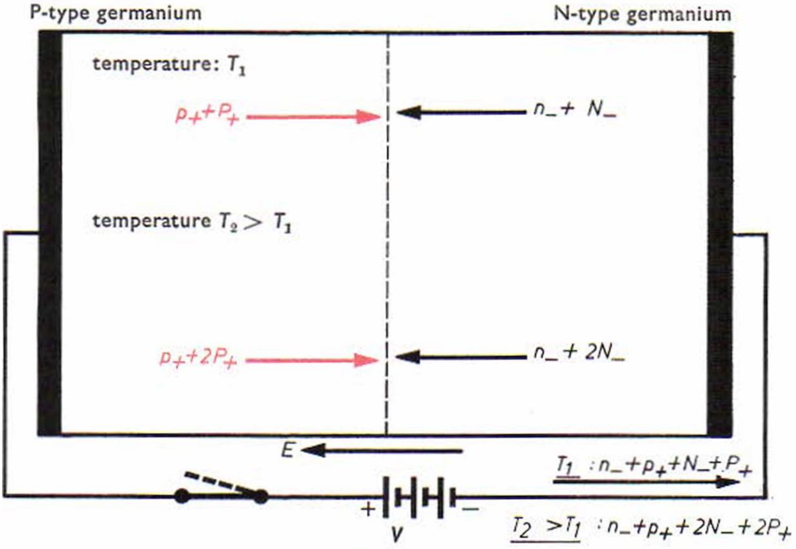

P-type germanium
N-type germanium
temperature: T_1
$p_+ + P_+$
$n_- + N_-$
temperature $T_2 > T_1$
$p_+ + 2P_+$
$n_- + 2N_-$
E
T_1 : $n_- + p_+ + N_- + P_+$
+
−
V
$T_2 > T_1$: $n_- + p_+ + 2N_- + 2P_+$

Fig. 63d

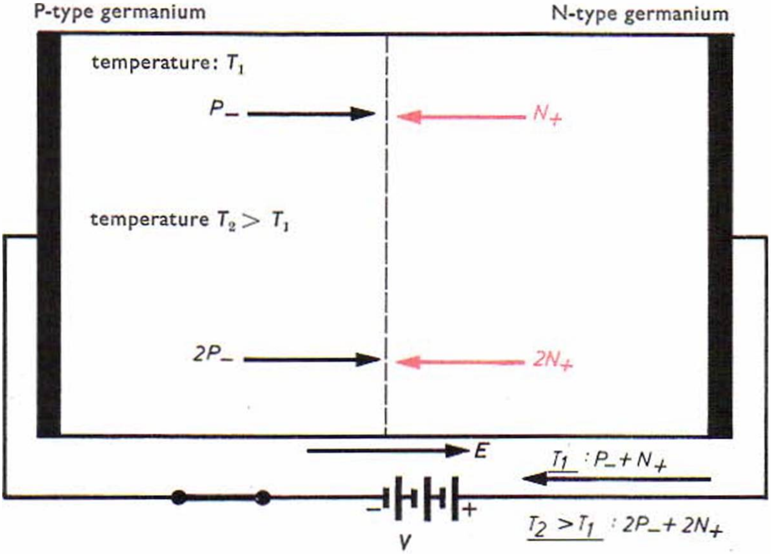

P-type germanium
N-type germanium
temperature: T_1
P_-
N_+
temperature $T_2 > T_1$
$2P_-$
$2N_+$
E
T_1 : $P_- + N_+$
−
+
V
$T_2 > T_1$: $2P_- + 2N_+$

Fig. 63e

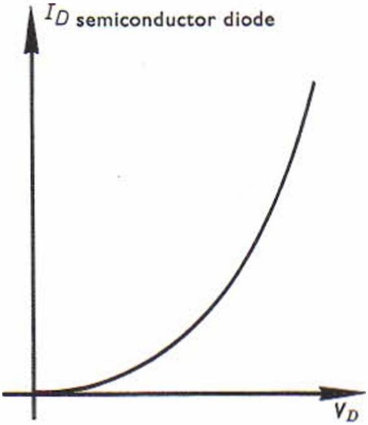


Fig. 64

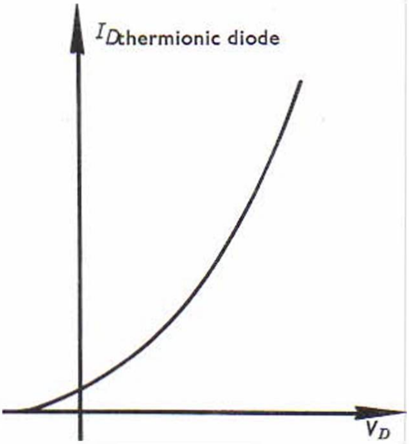


Fig. 65

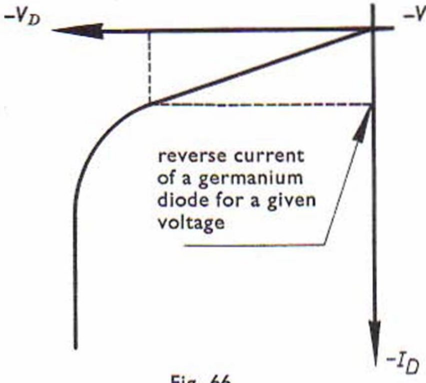


Fig. 66

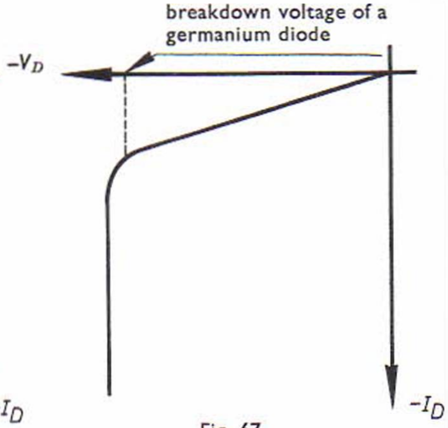


Fig. 67

CHAPTER 8

Comparison between thermionic diodes and semiconductor diodes

Semiconductor diodes offer the following advantages:

1) The absence of a heater. No heater current source is required, and this considerably reduces the problem of hum.
2) Very small capacitance (at least for point-contact diodes). The capacitance of a germanium diode of this type is about 1 pF; this advantage is particularly important in the use of these diodes at high frequencies.
3) The forward resistance of a semiconductor diode is lower than that of a thermionic diode.
4) In a closed circuit, the semiconductor diode does not produce any current in the absence of a voltage (Fig. 64), whereas a thermionic diode does produce a current (Fig. 65). In measuring-circuits, it is often necessary to compensate for this current.
5) Semiconductor diodes can easily be incorporated in the wiring.
6) Both their weight and volume are extremely small.

Because of these advantages the semiconductor diode is often preferred for numerous applications, but it must not be forgotten that it also has a number of disadvantages:

1) A thermionic diode does not pass any current at all in the reverse direction, while at high voltages a germanium diode passes a current which cannot be neglected (Fig. 66).
2) The thermionic diode can stand high reverse voltages, which is not true of the semiconductor diode, whose breakdown voltage ($-V_{DM}$) is not so high (Fig. 67).
3) The characteristics of the germanium diode change with the temperature, not only in the forward direction but particularly in the reverse direction.

CHAPTER 9

Rectification

The semiconductor can be used as a rectifying element in half-wave rectifiers, full-wave rectifiers, voltage-doubler circuits, and in bridge circuits etc.

9.1. Maximum permissible reverse voltage $-V_{DM}$

The amplitude of the alternating voltage which may be connected to the terminals of a semiconductor diode is limited by the reverse characteristic of the diode and also by the power which is dissipated in the junction. We will illustrate this for the OA 85 germanium diode.
At an ambient temperature of 25° C the maximum reverse voltage $-V_{DM}$ is 115 V. This voltage is the most important factor in designing a rectifier circuit incorporating a diode. We must distinguish between the following two cases:

a) Resistive load

This case is represented by the circuit of Fig. 68. We will assume that the input voltage is sinusoidal (Fig. 69). Fig. 70 shows the characteristic of the OA 85. A closer examination shows us the following sequence of events. At instant t_0 there is no voltage across the circuit and the working point of the characteristic is at *A*. At instant T_1 the anode is positive with respect to the cathode; the diode conducts, and the current increases from 0 to a maximum value, point *B*. At instant t_2 the voltage is again zero, and no current flows through the circuit. At instant t_3 the diode is connected in the reverse direction and a low reverse current flows through the circuit, point *D*. At instant t_4 the voltage is again zero and no current flows through the circuit, point *E*.
The reverse voltage is a maximum at instant t_3, and this voltage must not be allowed to exceed the breakdown voltage of the diode $-V_{DM}$. This means that the peak value of the applied voltage must be lower than the voltage $-V_{DM}$.

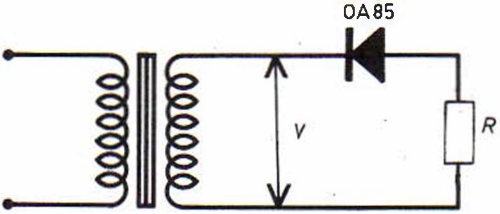


Fig. 68

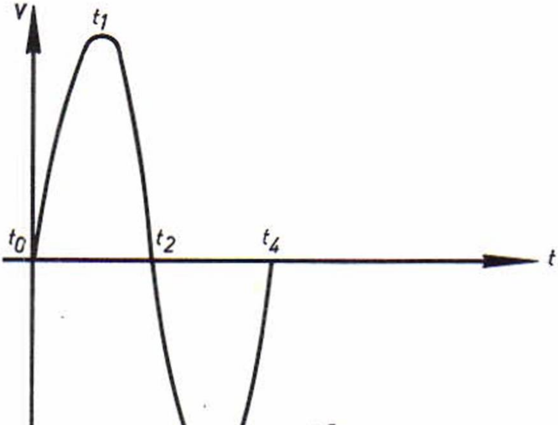


Fig. 69

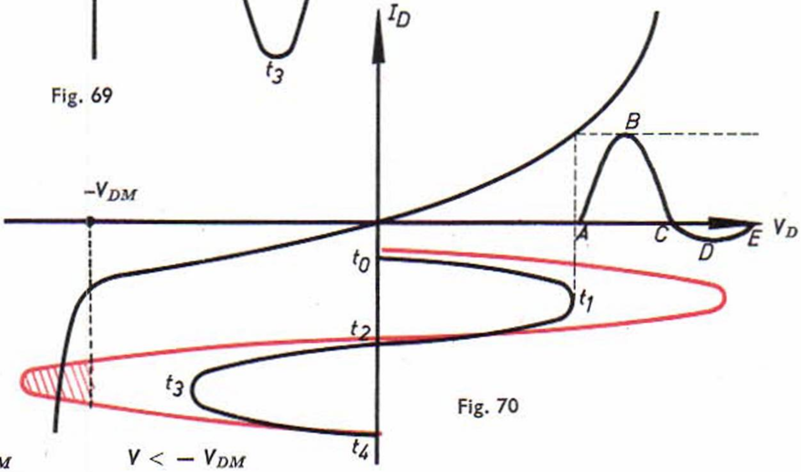


Fig. 70

$> - V_{DM}$ $V < - V_{DM}$

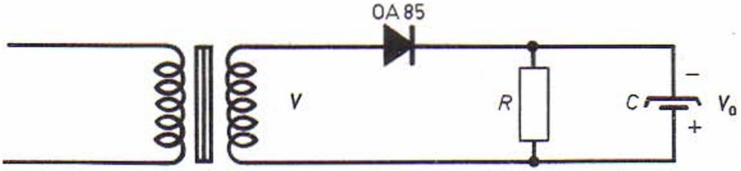


Fig. 71

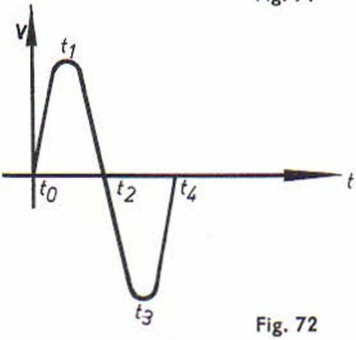


Fig. 72

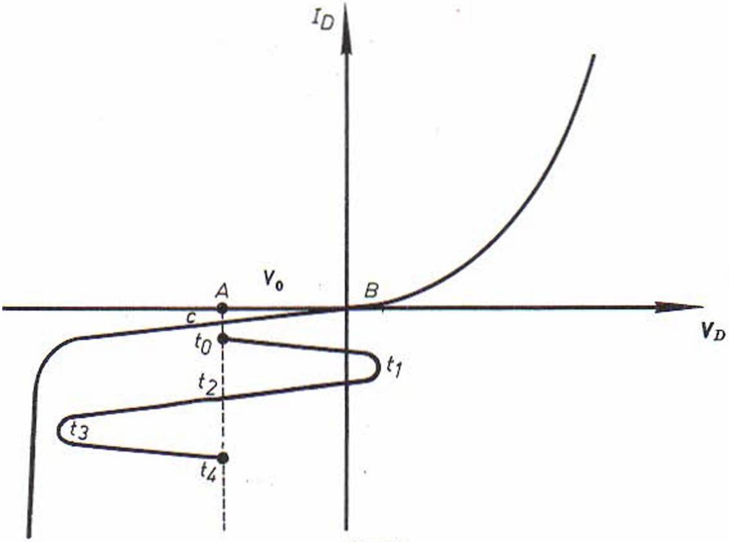


Fig. 73

For an OA 85 diode, the r.m.s. value of the maximum permissible voltage which may be applied to this circuit is: $115/\sqrt{2}$. ile approx. 82 V. This means that to rectify an alternating voltage having an r.m.s. value of 110 V, it would be necessary to connect two diodes of this type in series.

b) Resistive – capacitive load

In the circuit of Fig. 71, the load consists of a resistor R and a capacitor C. It will again be assumed that the applied voltage is absolutely sinusoidal (Fig. 72); the characteristic of the diode is shown once more in Fig. 73. When the diode conducts, the capacitor quickly becomes charged. On the other hand if the diode is blocked, the capacitor will slowly discharge through the resistor R, which constitutes the load of the circuit. The resistance of R will always be very large in comparison with the forward resistance of the diode, so that the capacitor will always remain charged to a considerable direct voltage $-V$. This means that the diode is biased, so that the alternating voltage is applied relative to this point $-V$ (Fig. 73). If we consider what is happening at various instants we find the following. At instant t_0 – at point A on the characteristic – the diode is blocked and the capacitor is slowly discharching through the resistor R. At instant t_1 a very small current is flowing through the diode – point B on the characteristic – and the capacitor is becoming charged. At instant t_2 the diode is blocked once more, and the capacitor begins to discharge through the resistor R (point C). At instant t_3 the voltage across the diode is equal to the sum of the voltage across the capacitor ($-V$) and the peak value of the applied voltage. The capacitor is still discharging slowly through the resistor R. At instant t_4 the diode is still blocked by the bias voltage across the capacitor.

The diode is found to be permanently biased, as a result of which the maximum permissible reverse voltage is also lower (Fig. 74). In those cases in which the time constant RC of the circuit is large in relation to one cycle of the input voltage, the r.m.s. value of the maximum permissible reverse voltage which may be applied to an OA 85 diode is:

$$115/2\sqrt{2} \approx 41 \text{ V}.$$

9.2. The value of the load resistance

As a general rule the load resistance must be small in relation to the reverse resistance of the diode. In the above case the capacitor discharges between instants t_2 and t_4 through both the load R and through the reverse resistance of the diode. Now the discharge current through the load R must be large in relation to the reverse current flowing through the diode. For this reason R will always be made smaller than the reverse resistance of the diode. For an OA 85 diode, R is not usually made any higher than 2kΩ.

9.3. Non-sinusoidal input voltage

It is also usual to specify a maximum average value for the input voltage. For the OA 85 diode, this voltage is $-V_D = 90$ V at a temperature of 25° C.

9.4. The rectified current

The rise in temperature of the diode which is produced by the rectified current sets a limit to the latter. The increase in temperature depends directly on the r.m.s. value of the current flowing through the diode.

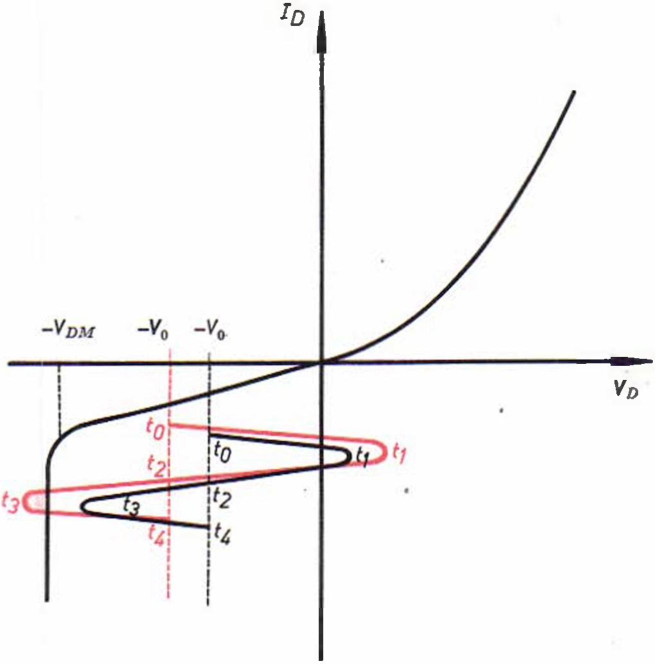

I_D
V_D
$-V_{DM}$
$-V_0$
$-V_0$
t_0
t_0
t_1
t_1
t_2
t_2
t_3
t_3
t_4
t_4

Fig. 74

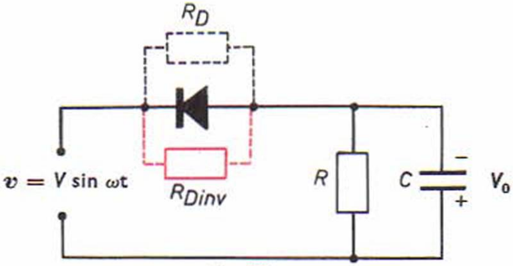

R_D
R_{Dinv}
$v = V \sin \omega t$
R
C
−
+
V_0

Fig. 75

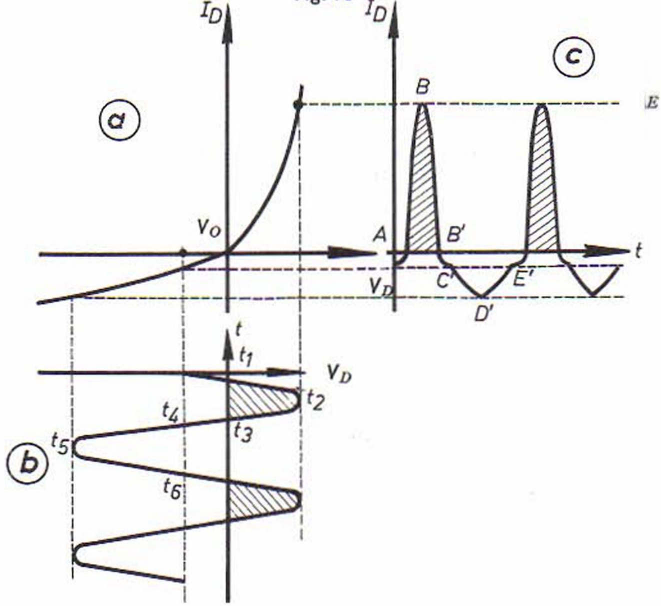

I_D
I_D
a
b
c
B
E
V_O
A
B'
t
V_D
C'
E'
D'
t
t_1
V_D
t_2
t_3
t_4
t_5
t_6

Fig. 76

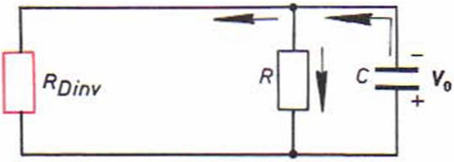


Fig. 77

CHAPTER 10

Comparison between the operation of a semiconductor diode and a thermionic diode as detector

The quality of a detector circuit is principally determined by the following two factors:

a) The detection efficiency.
b) The damping resistance.

We will investigate to what extent the specific characteristics of the two types of diode can affect these quantities. From this and from considerations concerning supply, dimensions, etc. we shall see in which cases one type of diode is to be preferred to another.

10.1. Detection efficiency

Fig. 75 shows a much-used detector circuit. The load on the diode consists of a resistor *R* and a capacitor *C*. The voltage on the input of the detector is supplied by a generator whose internal resistance we shall regard as being infinitely small; the amplitude of this voltage is constant. In Fig. 76, *a* is the diode characteristic, *b* is the input voltage as a function of time, and *c* is the current through the diode as a function of time.

We shall represent the peak value of the signal supplied by the generator by *E*. The time constant of the load circuit is very small; the capacitor charges up very quickly via the load resistance of the diode in the forward direction (interval *AB*, Fig. 76).

Fig. 77 shows the resistances across which the capacitor discharges, that is the load resistor *R* with the reverse resistance $R_{D\,\mathrm{inv}}$ of the diode parallel to it. From t_3 to t_4 the capacitor discharges slowly; the reverse current flowing through the diode is very small, because the reverse resistance of the diode is very large (*B'C'* Fig. 76c). From t_4 to t_5 the reverse voltage increases, as a result of which the reverse resistance decreases, so that the reverse current also increases (*C'D'* Fig. 76c). From t_5 to t_6 the reverse voltage decreases, while the reverse resistance increases; in this way the reverse current decreases, as also the discharging current of the capacitor (*D'E'* Fig. 76c).

The voltage V across the capacitor adjusts itself to a value such that the total charge conveyed to the capacitor during one cycle is equal to the total charge removed. The maximum value which this voltage V can reach is equal to the peak value E of the input signal. This value could only be reached if the reverse resistance of the diode $R_{D\,\mathrm{inv}}$ and the load resistance R were infinitely large. By definition, the detection efficiency is:

$$\eta = V_o/E.$$

In order to simplify the calculations we shall assume that the diode characteristic is ideal, and will take no account of the variations in the resistance of the diode in both forward and reverse directions which result from the applied voltage. The detection efficiency of such a germanium diode with an ideal characteristic (Fig. 78), could be calculated by equating the charge and the discharge of the capacitor.

It can be demonstrated that the detection efficiency is inversely proportional to:

$$(G + G_{D\mathrm{inv}})\,/G_D,$$

where $G = 1/R$ is the conductance of the load, $G_D = 1/R_D$ is the forward conductance of the diode, and $G_{D\mathrm{inv}} = 1/R_{D\mathrm{inv}}$ is the reverse conductance of the diode.

The curve represented in Fig. 79 shows that the efficiency is a maximum when this function is zero. The detection efficiency thus increases with decreasing values of G and $G_{D\mathrm{inv}}$ (that is, if the load resistance and the reverse resistance of the diode are very large) and if G_D is very large (that is, if the forward resistance is very small).

Now in contrast to a germanium diode, a thermionic diode has an infinite reverse resistance, so that it is easier to obtain a detection efficiency of almost 1 with this type of diode than it is with a germanium diode.

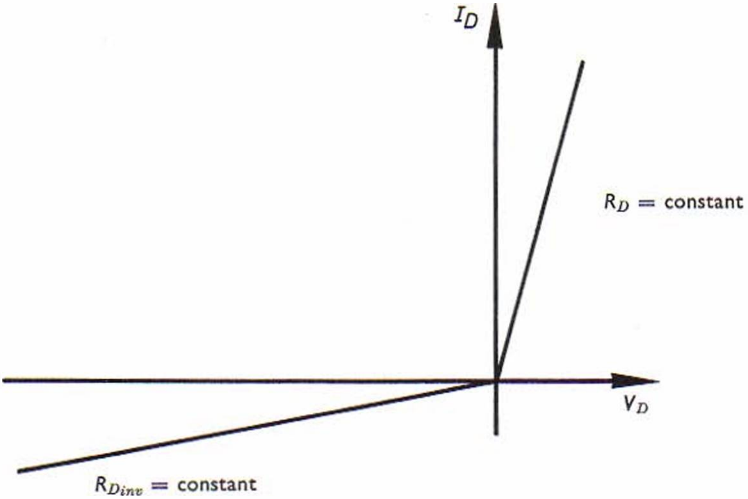

I_D
R_D = constant
V_D
R_{Dinv} = constant

Fig. 78

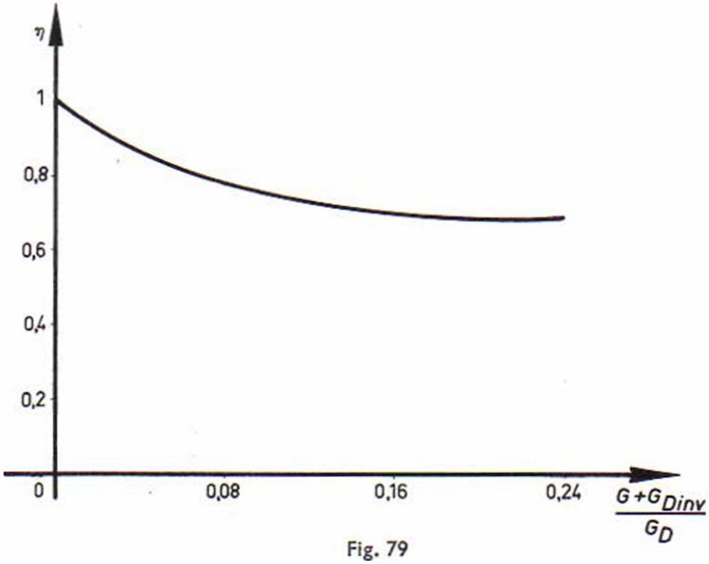
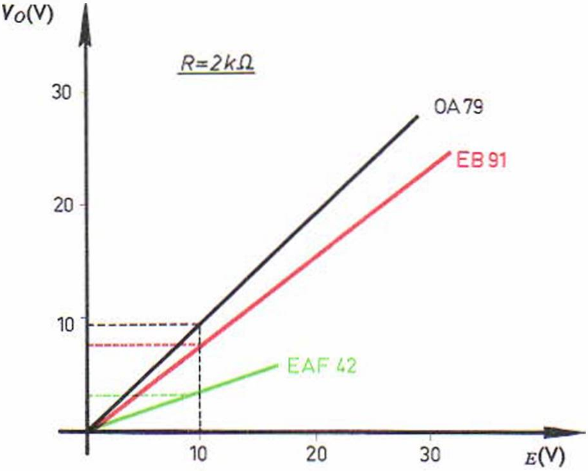
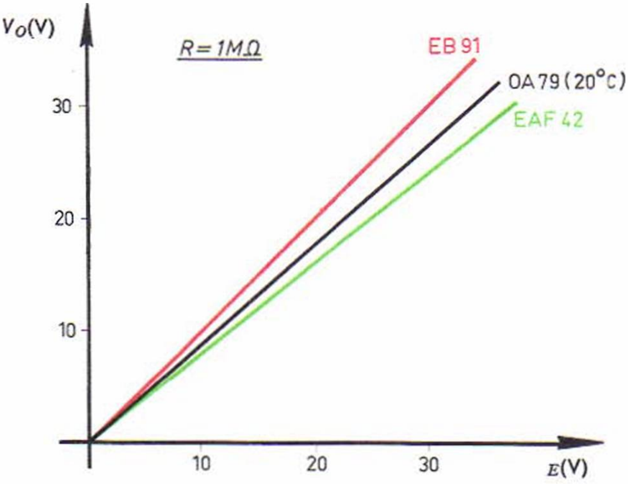

η
1
0,8
0,6
0,4
0,2
0
0,08
0,16
0,24
$\frac{G+G_{Dinv}}{G_D}$

Fig. 79



V_O(V)
R=2kΩ
30
20
10
OA 79
EB 91
EAF 42
10
20
30
E(V)

Fig. 80



V_O(V)
R=1MΩ
EB 91
OA 79 (20°C)
EAF 42
30
20
10
10
20
30
E(V)

Fig. 81

10.2. Determination of the optimum value of the load resistance for a semiconductor diode and for a thermionic diode

We can plot the direct voltage V which is built up across the capacitor C as a function of the peak value E of the applied signal. In Fig. 80 this has been done for a low value of the load resistance, namely for $R = 2\text{k}\Omega$. Curves have been plotted for a germanium diode (OA 79) and for two thermionic diodes (EB 91 and EAF 42). Fig. 81 is a similar graph drawn for a high load resistance, i.e. for $R = 1\text{M}\Omega$.

In all these cases the diode characteristic has been taken as ideal, that is, it is assumed that both the forward resistance and the reverse resistance are independent of the amplitude of the input voltage. As already explained, however, the input voltage does have a considerable effect on these resistances.

10.3. Low load resistance *R*

If the load resistance R has a low value, the highest efficiency is obtained by using a germanium diode. For a signal of amplitude $E = 10$ V, the direct voltage V across the capacitor is;

approx. 9.9 V for an OA 79
approx. 8.5 V for an EB 91 and
approx. 3 V for an EAF 42.

The highest detection efficiency is thus obtained by using a germanium diode.

Explanation of the curves

As already explained, the detection efficiency is inversely proportional to the quotient:

$$(G + G_{D\,\text{inv}}) / G_D.$$

We will now determine the values of this quotient for a germanium diode (OA 79) and for a thermionic diode (EB 91).

a) Germanium diode (OA 79)

Suppose that $R_{D\,\mathrm{inv}} = 1\mathrm{M}\Omega$, $R_D = 200\Omega$, $R = 2000\Omega$. In this case:

$$G = 1/R = 1/2000 = 5 \times 10^{-4}\ \mathrm{A/V} = 500\mu\mathrm{A/V},$$

$$G_{D\mathrm{inv}} = 1/R_{D\mathrm{inv}} = 1/10^{6} = 10^{-6}\ \mathrm{A/V} = 1\ \mu\mathrm{A/V}$$

and

$$G_D = 1/R_D = 1/2000 = 5 \times 10^{-3}\ \mathrm{A/V} = 5000\ \mu\mathrm{A/V}.$$

From this follows:

$$(G + G_{D\mathrm{inv}})/\ G_D = (500 + 1)/5000 \approx 0.1.$$

b) Thermionic diode (EB 91)

Suppose that $R_{D\mathrm{inv}} = \infty$, $R = 2000\Omega$, $R_D = 300\Omega$. In this case:

$$G = 1/R = 500\ \mu\mathrm{A/V},$$

$$G_{D\mathrm{inv}} = 1/R_{D\mathrm{inv}} = 1/\infty = 0, \text{ and}$$

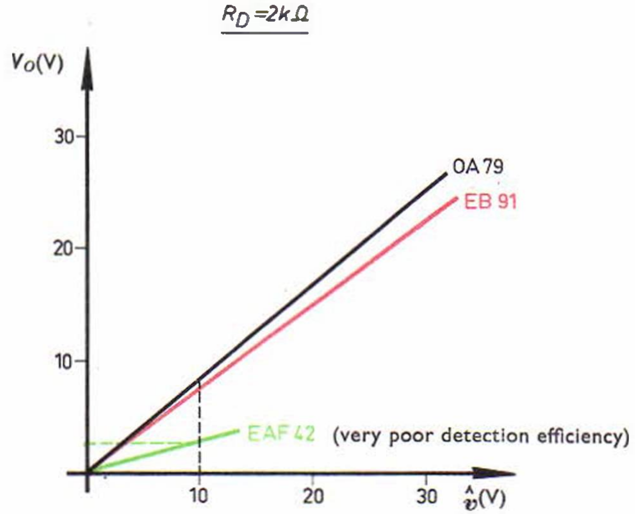
$$G_D = 1/R_D = 1/300 \approx 3 \times 10^{-3}\ \mathrm{A/V} = 3000\ \mu/\mathrm{AV}.$$

From which follows:

$$(G + G_{D\mathrm{inv}})/G_D = (500 + 0)/3000 \approx 0.17.$$

For the germanium diode we can neglect $G_{D\mathrm{inv}}$ in relation to G, and the quotient $(G + G_{D\mathrm{inv}})/G_D$ is approx. 0.1, while for a thermionic diode, this quotient is approx. 0.17.

We see that with a load resistance having a low value, the effect of the reverse conductance $G_{D\,\mathrm{inv}}$ can be neglected relative to the load conductance G.



$R_D = 2k\Omega$
V_O(V)
30
20
10
OA 79
EB 91
EAF 42
(very poor detection efficiency)
10
20
30
$\hat{v}$(V)

Fig. 82

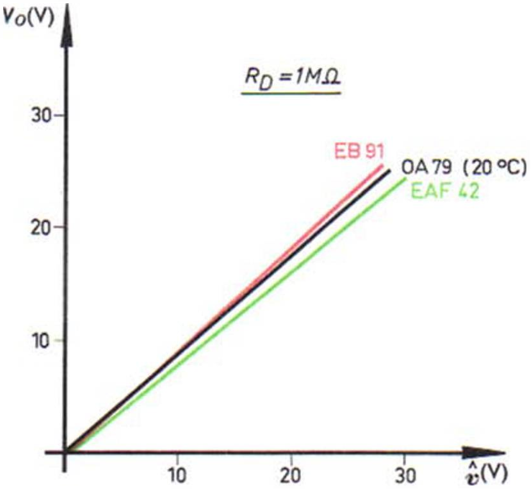


Fig. 83

As the forward resistance of the germanium diode is lower than that of a thermionic diode, its conductance G_D is greater than that of a thermionic diode. The ratio G/G_D is consequently smaller for a germanium diode, than for a thermionic diode. From this we see that for a low load resistance a higher detection efficiency can be obtained with a germanium diode than with a thermionic diode.

10.4. High load resistance R

Fig. 83 shows that for $R = 1\text{M}\Omega$ a higher efficiency can be obtained with an EB 91 thermionic diode than with an OA 79 germanium diode.

a) Germanium diode (OA 79)

Suppose that $R_{D\text{inv}} = 1\text{M}\Omega$, $R_D = 200\Omega$, $R = 1\text{M}\Omega$. In this case:

$$G = 1/10^6 = 10^{-6}\ \text{A/V} = 1\ \mu\text{A/V},$$

$$G_{D\text{inv}} = 1/10^6 = 10^{-6}\ \text{A/V} = 1\ \mu\text{A/V, and}$$

$$G_D = 1/200 = 5 \times 10^{-3}\ \text{A/V} = 5000\ \mu\text{A/V}.$$

From which follows:

$$(G + G_{D\text{inv}})/G_D = (1 + 1)/5000 = 0.0004.$$

b) Thermionic diode (EB 91)

Suppose that $R_{D\text{inv}} = \infty$, $R_D = 300\Omega$, $R = 1\text{M}\Omega$. In this case:

$$G = 1/10^6 = 10^{-6}\ \text{A/V} = 1\ \mu\text{A/V},$$

$$G_{D\text{inv}} = 1/\infty = 0\text{, and}$$

$$G_D = 1/300 \approx 3 \times 10^{-3}\ \text{A/V} = 3000\ \mu\text{A/V}.$$

From which follows:

$$(G + G_{D\text{inv}})/G_D = (1 + 0)/3000 \approx 0.0003.$$

Thus we see that for the germanium diode the ratio $(G + G_{D\text{inv}})/G_D$ is 0.0004 (at a temperature of 20° C) and for the thermionic diode it is 0.0003.

With a high load resistance, it is evident that the effect of the reverse resistance of a germanium diode may no longer be neglected, and this effect also reduces the detection efficiency of a germanium diode relative to that of a thermionic diode. With a high load resistance, a higher detection efficiency is obtained with a thermionic diode than with a germanium diode.

10.5. The effect of temperature on the detection efficiency of a germanium diode

a) Low load resistance

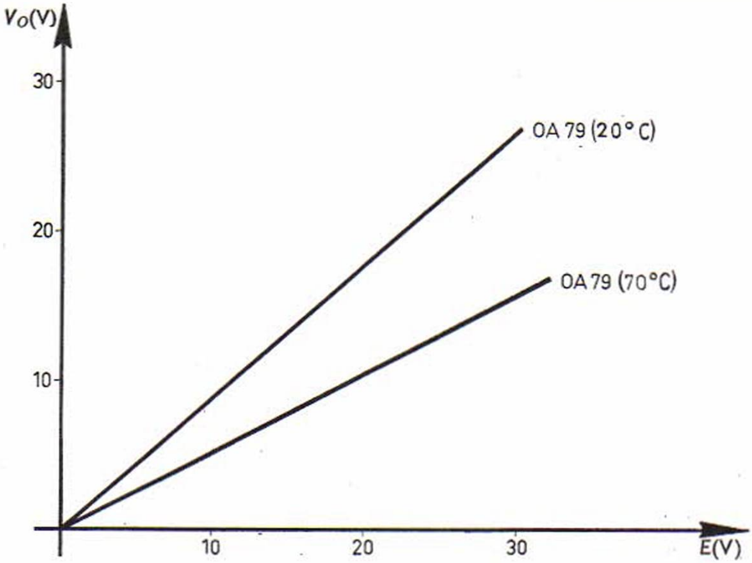
If the load resistance has a low value, the influence of the reverse resistance of the diode may be neglected. It goes without saying that the effect of changes in this resistance can most certainly be neglected. Consequently the temperature will not have any effect worth mentioning on the detection efficiency if the load resistance is low.

b) High load resistance

A high load resistance means a low value of the conductance G, so that the latter will have little effect on the ratio $(G + G_{D\mathrm{inv}})/G_D$, which is thus mainly determined by the conductance $G_{D\mathrm{inv}}$ of the diode in the reverse direction. If the temperature increases, this conductance will also increase and consequently the detection efficiency of the germanium diode will decrease (see Fig. 84).

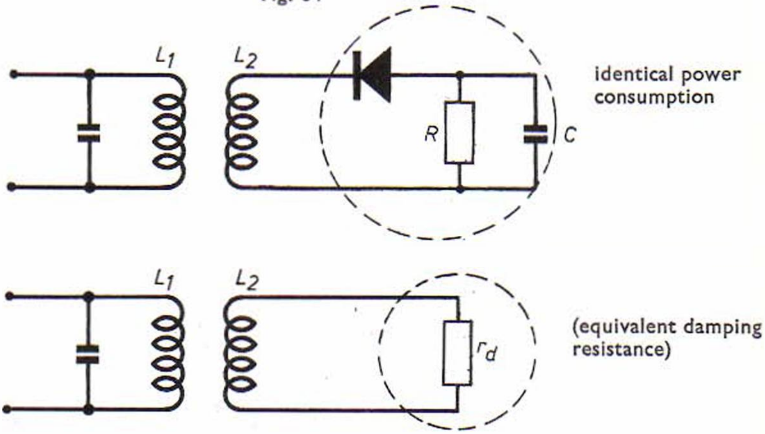
10.6. Damping resistance

In addition to the detection efficiency, the damping which the detector circuit exercises on the previous tuned circuit is an important factor. This damping depends both on the detector circuit itself, and on the form and amplitude of the input signal. By damping resistance we mean that resistance which could replace the detector circuit, and cause the same amount of damping. We will represent this resistance by the symbol r_d (Fig. 85). If the output power is P, and the peak value of the input signal is E, the equivalent resistance is $r_d = E^2/2P$. In this connection, we will re-examine the circuit of Fig. 75. If the load resistance, the capacitor and the reverse resistance of the diode all have high values, the direct voltage V across the capacitor will be near enough equal to the peak voltage E of the input signal. The power dissipated is then $P = E^2/R$. The equivalent resistance of a thermionic diode (whose reverse resistance is infinite) is equal to half the value of the load resistance, thus the damping resistance



V_O(V)
30
20
10
10
20
30
E(V)
OA 79 (20° C)
OA 79 (70°C)

Fig. 84



L_1
L_2
R
C
identical power consumption
L_1
L_2
r_d
(equivalent damping resistance)

Fig. 85

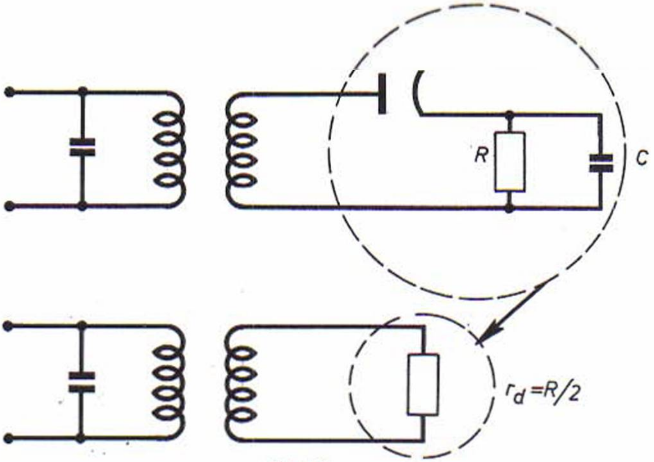


Fig. 86

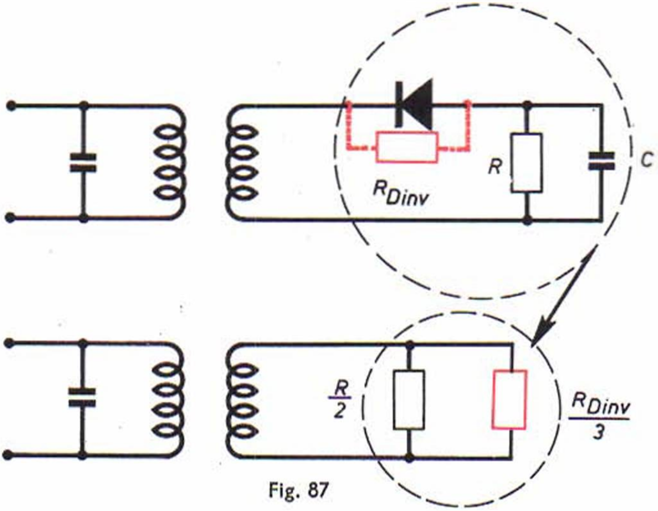

R
C
R_{Dinv}
$\frac{R}{2}$
$\frac{R_{Dinv}}{3}$

Fig. 87

$$r_d = R/2.$$

By contrast, the equivalent resistance of a germanium diode is equal to half the load resistance connected in parallel with one third of the reverse resistance of the diode (see Fig. 87):

$$1/r_d = 2/R + 3/R_{D\mathrm{inv}}.$$

If the temperature of a germanium diode or the input voltage increases, the reverse resistance of the diode decreases. Consequently the damping resistance also decreases with increasing temperature or input voltage, at least if the load resistance has a high value.

With decreasing values of the load resistance, the attenuation which is caused by the reverse resistance of the diode will decrease as in this case $2/R$ becomes large in relation to $3/R_{D\mathrm{inv}}$. In practice, the attenuation which must be ascribed to the reverse resistance of a germanium diode becomes negligible when the load resistance is of the order of a few kilohms. The damping resistance depends particularly on the load resistance and on the reverse resistance of the diode. Summing up, the damping resistance depends on:

- the characteristics of the diode,
- the ambient temperature of the diode,
- the amplitude of the input signal

CHAPTER 11

Operation at radio frequencies

The efficiency of a germanium diode when used as a detector at radio frequencies (for example, of the order of 40 Mc/s) cannot be deduced from the static characteristics.

Suppose that an OA 73 germanium diode is used as video detector with a load resistance of 3.9 kΩ which is connected in parallel with a 20 pF capacitor (see Fig. 88). The detection efficiency of this circuit is found to be 0.55 at a frequency of 20 Mc/s and 0.53 at a frequency of 40 Mc/s. From this it follows that the detection efficiency at these frequencies is lower than it is at very low frequencies. This is principally due to the fact that the time constant of the load circuit is no longer very large in relation to the duration of one cycle of the signal. In addition, the capacitance of the diode and the load capacitance *C* form a capacitive voltage-divider, which reduces the input signal to the diode (Fig. 89).

For the two frequencies which have been mentioned, the difference in detection efficiency is only small, but the difference in damping on the previous tuned circuit proves to be very great:

At a frequency of 20 Mc/s (Fig. 90) $r_d=4500\Omega$
At a frequency of 40 Mc/s (Fig. 91) $r_d=3000\Omega$

While the direct voltages across the load resistance are thus practically identical in both cases, the damping exercised by the detector circuit on the previous stage proves to be considerably greater at 40 Mc/s, so that the amplification of the last r.f. stage or i.f. stage will be rather lower.

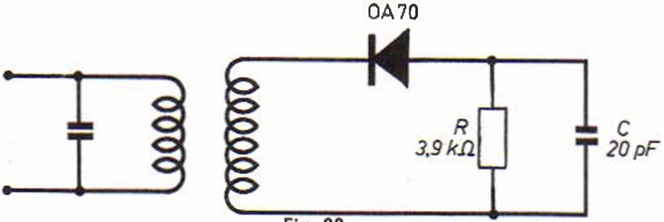

OA70
R
3,9 kΩ
C
20 pF

Fig. 88

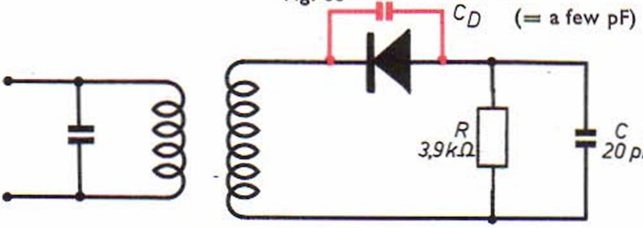

C_D
(= a few pF)
R
3,9 kΩ
C
20 pF

Fig. 89

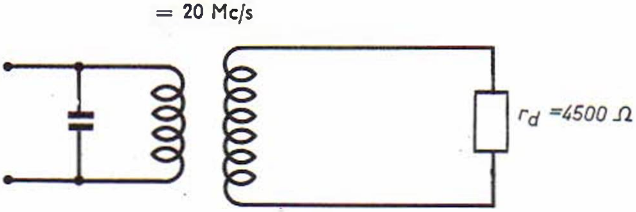

= 20 Mc/s
r_d =4500 Ω

Fig. 90

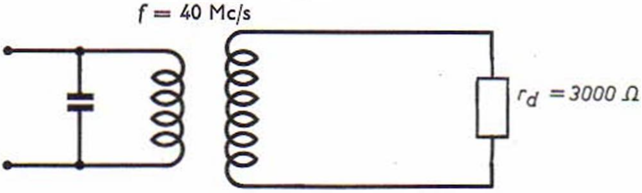

f = 40 Mc/s
r_d = 3000 Ω

Fig. 91

TYPE	CONSTRUCTION	APPLICATION AND CHARACTERISTICS
OA5	Gold-bonded germanium diode for single-ended connection.	Switching diode; low forward resistance.
OA7	Gold-bonded germanium diode for single-ended connection.	Rapid switching; low forward resistance.
OA9	Gold-bonded germanium diode for single-ended connection.	Rapid switching; very low forward resistance.
OA70	Germanium point-contact diode.	Video detector.
OA73	Germanium point-contact diode.	Video detector.
OA79	Germanium point-contact diode.	Ratio detector
2–OA79	Matched pair of OA79 diodes.	Ratio detector
OA81	Germanium point-contact diode.	High reverse voltage
OA85	Germanium point-contact diode.	High reverse voltage and low forward resistance.
OA86	Germanium point-contact diode.	Rapid switching: high reverse voltage.
OA47	Gold-bonded germanium diode for double-ended connection; subminiature.	as type OA7.
OA90	Germanium point-contact diode; subminiature.	as type OA70.
OA91	Germanium point-contact diode; subminiature	as type OA81
OA92	Germanium point-contact diode; subminiature.	Rapid switching; low reverse voltage.
OA95	Germanium point-contact diode; subminiature.	as type OA85.
AAY11	Germanium point-contact diode; subminiature.	as type OA86.
BA100	Silicon diode: subminiature	General purpose diode for television receivers.
OA200	Silicon diode: subminiature	General purpose.
OA202	Silicon diode: subminiature	General purpose.
BA102	Alloyed silicon diode; subminiature	Automatic frequency control; voltagedependent capacitance.
OA31	Germanium junction diode	Supply units up to 12 A.
OA210	Alloyed silicon diode.	Supply circuits in television receivers for 110V mains voltage.
OA211	Alloyed silicon diode.	Supply circuits in television receivers for 220V mains voltage.
OA214	Alloyed silicon diode.	Supply circuits in television receivers for 250 V mains voltage.
BY100	Diffused silicon diode	Supply circuits in television receivers; permissible reverse voltage 800 V.
BYZ14	Diffused silicon diode	Supply units up to 20 A.

PART THREE

Transistors

CHAPTER 12

General considerations

The operation of transistors is also based on the physical properties of semiconductors. The transistor can be regarded as a combination of non-linear resistances whose values vary with the applied voltages and the polarity of these voltages, with the power dissipation, and with the ambient temperature. However, the transistor does not behave as a completely passive element. In the circuits in which it is used it is possible to obtain power amplification.

The application of the transistor is more complicated than that of the thermionic valve. As we shall explain, we have to deal not only with the familiar parameters (load resistance, voltage amplification), but also with new quantities (input resistance, current gain, internal feedback).

The most important source of data for the application of transistors consists of the various characteristic curves which are supplied by the manufacturer. The operation of the transistor – as passive element and as active element – can be examined with the aid of these characteristics. Originally, only point-contact transistors were available on the market, but these have now been completely replaced by junction transistors.

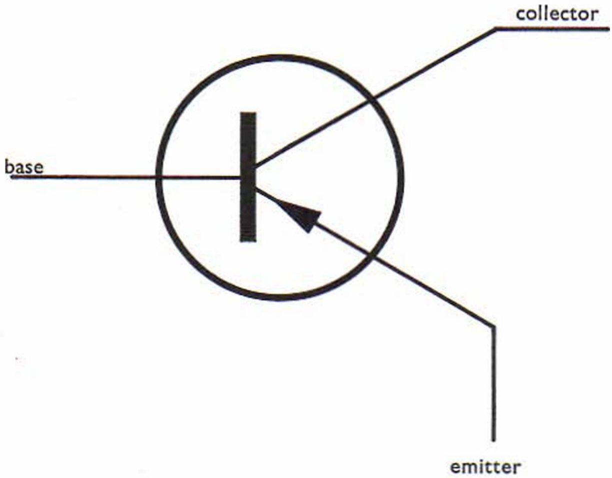

collector
base
emitter

Fig. 92

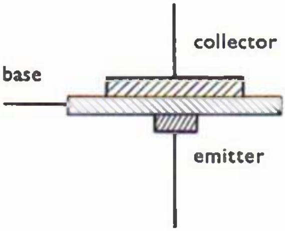

collector
base
emitter

Fig. 93

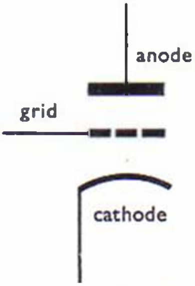

anode
grid
cathode

Fig. 94

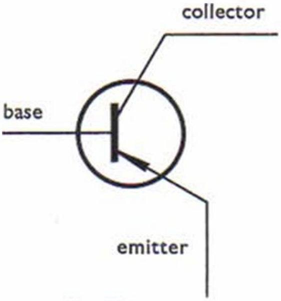

collector
base
emitter

Fig. 95

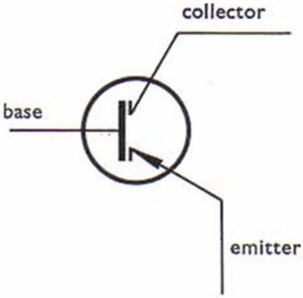

collector
base
emitter

Fig. 96

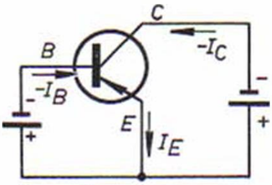


Fig. 97

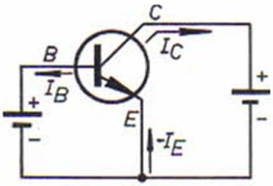


Fig. 98

CHAPTER 13

The junction transistor

Junction transistors consist of three pieces of germanium or silicon which constitute the three electrodes: the emitter, the base and the collector (see Fig. 93). To a certain extent these correspond respectively to the cathode, the grid and the anode of a triode (Fig. 94). In a junction transistor, the emitter and collector are always of the same type of germanium or silicon (usually *P*-type germanium) while the base is of the other type, (usually *N*-type germanium).

13.1. Circuit symbols

The symbol shown in Fig. 95 dates from the time when point-contact transistors were in use, and represents the two point-contacts resting on the base. This symbol is still normally employed for the junction transistor. Some publications employ the slightly different symbol of Fig. 96, which indicates that the emitter and the collector also consist of a layer of material. A distinction is made between *PNP*-junction transistors (Fig. 97) and *NPN*-junction transistors (Fig. 98).

13.2. Conduction in a transistor

In a thermionic valve, the current results from the movement of free electrons which are emitted by the cathode and attracted towards the anode. The origin of the current which flows in a transistor will be explained in a later chapter, but it is important to note at this stage that the directions of the currents flowing through a *PNP* transistor (Fig. 97) are opposite to the directions of those flowing an *NPN* transistor (Fig. 98). The voltages applied to the collector and base of a transistor always have the same polarity with respect to the emitter, being negative in relation to the emitter of a *PNP* transistor (Fig. 97), and positive in relation to the emitter of an *NPN* transistor (Fig. 98).

CHAPTER 14

The technology of the transistor

Transistors can be made by widely differing methods, but we will limit ourselves here to the general principles underlying the various operations involved in the manufacture of a transistor. The fundamental problem is that of obtaining pure germanium or silicon.

14.1. Pure germanium or silicon

The metal is first purified chemically. However, this purification is insufficient, and the metal must then be subjected to a special physical treatment which is based on the fact that impurities are more soluble in liquid germanium or silicon than in the solid.

For this treatment, the metal is placed in a quartz "boat" (Fig. 99), which moves along relative to a source of heat (a coil through which flows an r.f. current). The heat melts the metal in the zone which is exposed to the r.f. field. With the movement of the boat, the molten zone moves along the germanium or silicon. The metal which solidifies on the right hand side of the molten zone is purer than that which has not yet melted, on the left hand side of this zone. This operation is repeated several times, after which the impurities will be concentrated in the left hand end of the bar, which is then removed.

The germanium or silicon obtained in this way still contains an extremely small percentage of impurities (of the order of 10^{-8} to 10^{-9}). As a general rule, it is not possible to use pure metal for the manufacture of transistors; its characteristics must be changed by adding to it an accurately determined amount of impurity. This is done as follows.

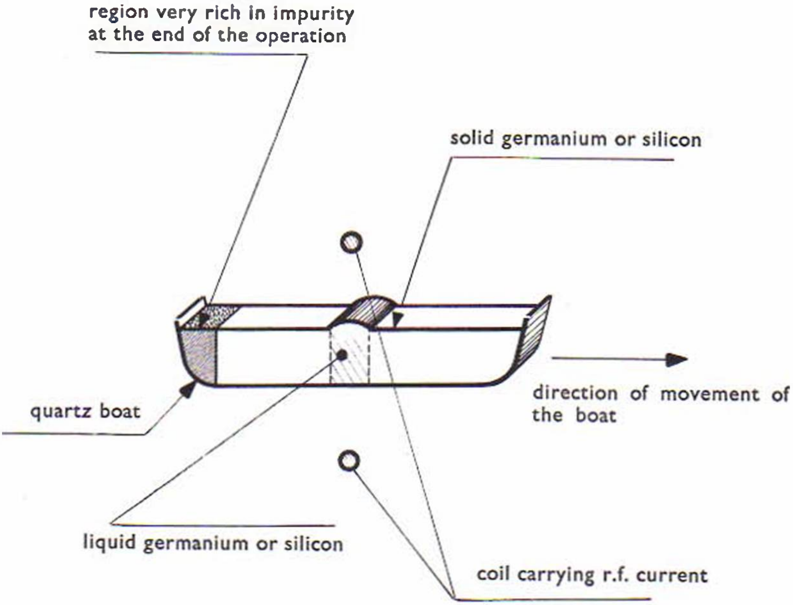

region very rich in impurity at the end of the operation
solid germanium or silicon
quartz boat
direction of movement of the boat
liquid germanium or silicon
coil carrying r.f. current

Fig. 99

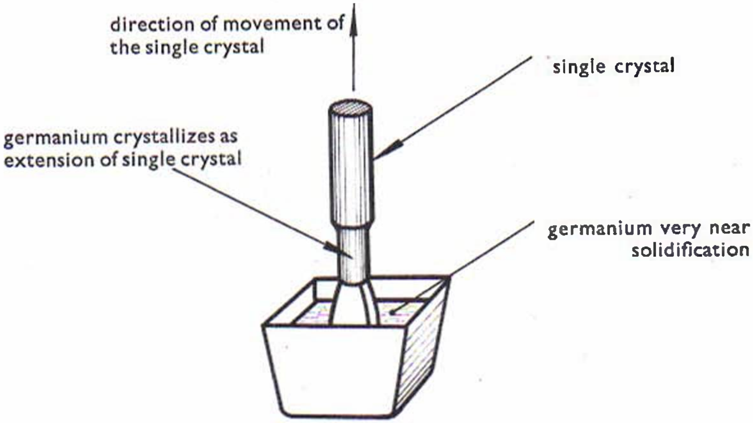

direction of movement of the single crystal
single crystal
germanium crystallizes as extension of single crystal
germanium very near solidification

Fig. 100

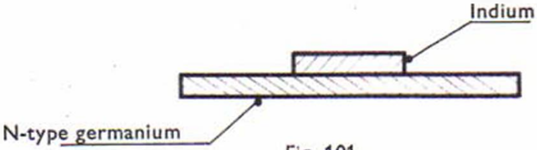

Indium
N-type germanium

Fig. 101

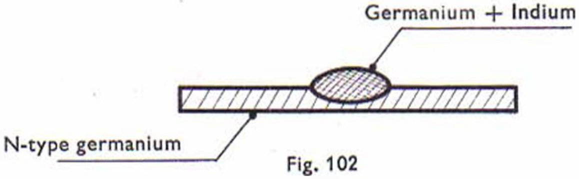

Germanium + Indium
N-type germanium

Fig. 102

14.2. The manufacture of *N*-type germanium

Purified germanium is fused in a quartz crucible after which the required amount of impurity is added. The germanium is then allowed to cool to a temperature which is only slightly above the melting point. A single crystal of germanium is now brought into contact with the surface of the liquid, and is raised extremely slowly. The germanium in contact with the single crystal also crystallizes in the single crystal form. In this way the crystal is continuously drawn up from the surface of the liquid until the crucible is empty. (Fig. 100). The percentage of impurity is now checked by measuring the resistivity of the crystal.

To make transistors such as the OC 41, the single crystal of *N*-type germanium which is obtained in this way is sawn into wafers, on both sides of which a layer of *P*-type germanium must be applied.

14.3. The manufacture of junction transistors

A small piece of indium is placed in contact with a wafer of germanium (Fig. 101) and the whole is placed in an oven in which the temperature is gradually increased to 500° C or 600° C. The indium melts first, (at a temperature of about 100° C) and germanium then dissolves in it until it is saturated. During cooling, germanium saturated with indium crystallises out, as a result of which *P*-type germanium is obtained, and a *PN*-junction is formed.

A similar junction can also be produced on the other side of the germanium wafer, and in fact the two operations can be carried out simultaneously.

CHAPTER 15

The mode of operation of thermionic valves

We will explain the mode of operation of thermionic valves with reference to Fig. 103. The electron current which flows through a valve is a function of the voltage between the grid and the cathode. The current increases as this negative voltage is decreased. Variations of the anode current are caused exclusively by variations of the grid voltage; no current flows between the grid and the cathode of the tube.

Fig. 104 represents the $I_a = f(V_g)$ characteristic of the valve. From this we see that the anode current is a maximum when no bias voltage is applied ($V_g = 0$), and increases no further if V_g is made positive.

15.1. Advantages

Valves can be used at very high frequencies.

Valves can supply higher powers.

The input impedance of valves is not limited.

The characteristics of valves are not affected by changes in temperature.

15.2. Limitations

Valves must have a heated cathode, which is an unavoidable source of hum.

Electric power has to be supplied to heat the heater.

All valves give rise to microphony to a greater or lesser extent.

The power efficiency of valves is fairly low, because of:

- their heater power.
- the high voltages which have to be applied in order to obtain a steep slope.

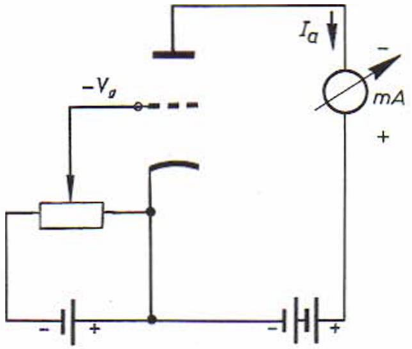


Fig. 103

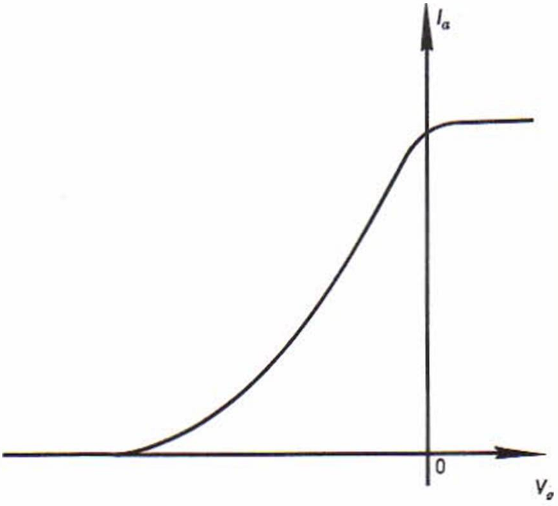


Fig. 104

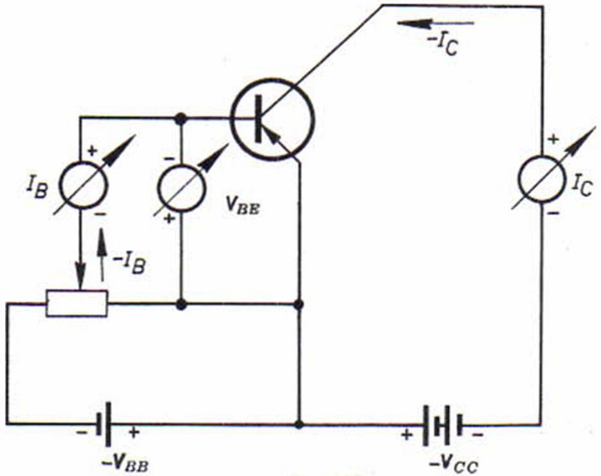

$-I_C$
I_B
+
−
$-I_B$
−
+
V_{BE}
+
−
I_C
−
+
$-V_{BB}$
+
−
$-V_{CC}$

Fig. 105

CHAPTER 16

The mode of operation of transistors

Fig. 105 represents a typical circuit. The electron current I_C flowing through the transistor is a function of the base-emitter voltage and of the current flowing in the base circuit. We will deal later with the characteristics which show the collector current as a function of the applied voltages and of the base current. The reader should note that, in contrast to thermionic valves, in which the anode current is a maximum if the grid and the cathode have the same potentials, the collector current of a transistor is a minimum if the base and the emitter have the same potential.

16.1. Advantages

Transistors operate at low voltages.
Transistors have a high power efficiency.
Transistors do not give rise to microphony.
The small size of transistors make it possible to keep the wiring compact, and simplifies the problem of screening.
The long life of transistors means that it is safe to solder them into wiring, thus preventing trouble due to bad contacts.

16.2. Limitations

The characteristics of transistors are affected by temperature changes.
The output power of transistors is limited.
There is a limit to the input impedance of transistors.

CHAPTER 17

PNP transistors

We start our consideration of *PNP* transistors with a thin block of *N*-type germanium; two regions of *P*-type germanium are applied, one on each side of this wafer, in such a way that the three regions form a single crystal (Fig. 106).

In the *P*-type germanium the impurity atoms are negatively ionized, and conduction is evidently due to holes. These move in the opposite direction to that of the applied electric field. In the *N*-type germanium the impurity atoms are positively ionized and conduction is due to electrons, which move in the direction of the applied electric field. In Fig. 106, *A*, *B* and *C* represent the three regions of this *PNP* material. Regions *A* and *C* are of *P*-type germanium, and region *B* is of *N*-type germanium. The boundaries seperating these three regions are indicated by broken lines.

In *PNP* transistors the percentage of impurity in the *P* regions is about 100 times the percentage of impurity in the *N* region, so that the hole density in regions *A* and *C* is very much greater than the free-electron density in region *B*.

17.1. Formation of the *PN*-junction between regions *A* and *B*

Region *A*, consisting of *P*-type germanium, is very rich in holes, while region *B*, consisting of *N*-type germanium, possesses few free electrons (see Fig. 107). The free electrons from region *B* diffuse into region *A*, where they will quickly fill up some of the very numerous holes in this region. The holes in region *A* diffuse into region *B*, where they will be filled up by the free electrons present (which are the majority carriers here, because they exceed the number of holes present).

In the neighbourhood of the junction there will be no free electrons left in region *B* and no mobile holes in region *A*. In region *A* only negatively ionized impurity atoms will remain, and in region *B* only positively ionized impurity atoms. In this way, an internal electric field of amplitude e_1 is produced, acting in the direction from the negatively ionized atoms to the positively ionized atoms. We will indicate the junction formed in this way by B_1.

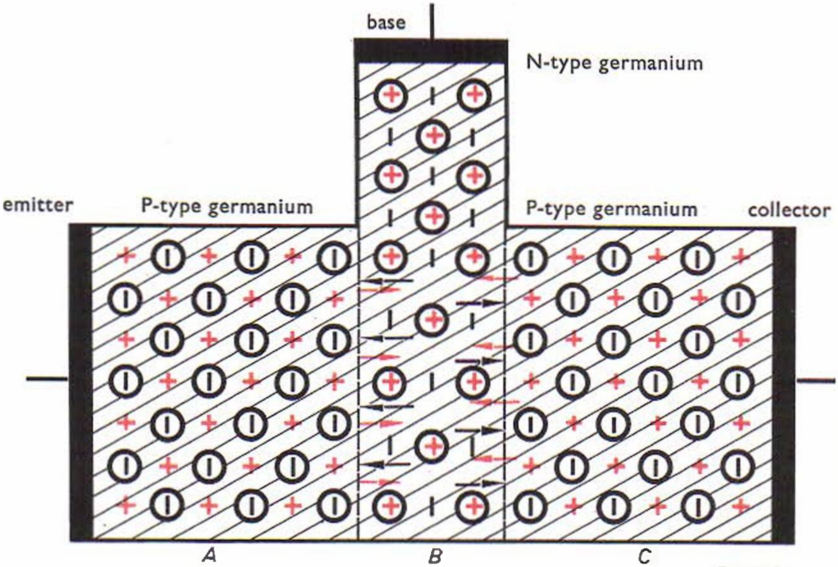

base
N-type germanium
emitter
P-type germanium
P-type germanium
collector
A
B
C

Fig. 106

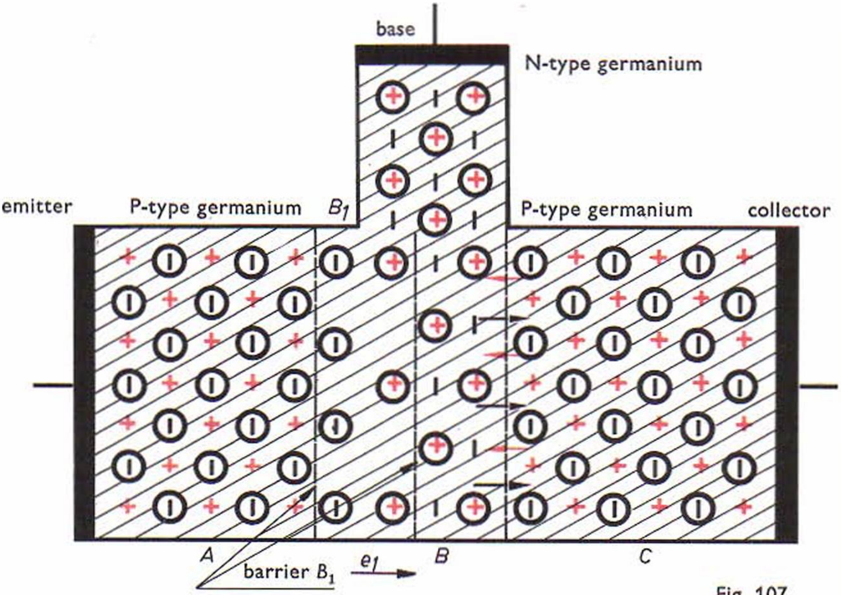

base
N-type germanium
emitter
P-type germanium
B_1
P-type germanium
collector
A
B
C
barrier B_1
e_1

Fig. 107

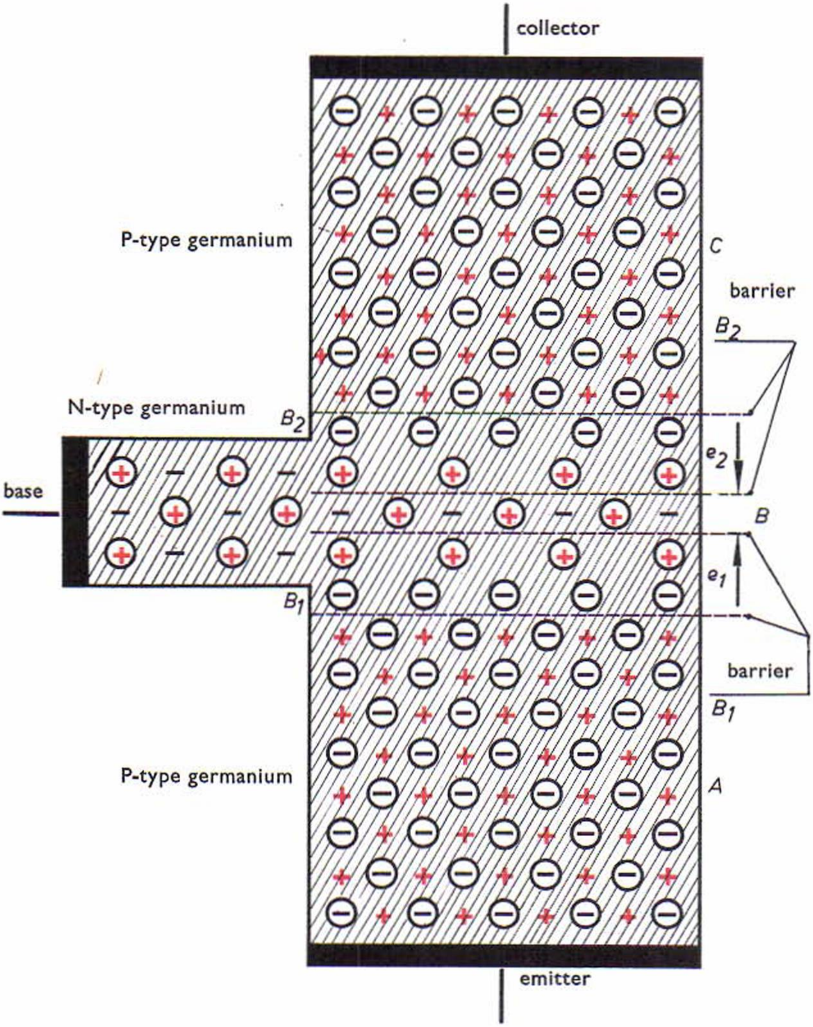

collector
P-type germanium
C
barrier
B_2
N-type germanium
B_2
e_2
base
B
e_1
B_1
barrier
B_1
P-type germanium
A
emitter

Fig. 108

17.2. Formation of the *NP*-junction between regions *B* and *C*

We will explain the formation of the junction between regions *B* and *C* with reference to Fig. 108. Region *B* consists of *N*-type germanium, which is rich in free electrons, and region *C* consists of *P*-type germanium, which is very rich in mobile holes. The free electrons from region *B* diffuse into region *C* and will fill up the numerous mobile holes in this region. On the other hand, the mobile holes from region *C* will diffuse into region *B*, and will very rapidly be filled up by the majority carriers in this region, that is the free electrons. In the neighbourhood of the boundary between *B* and *C*, there will no longer be any free electrons in region *B*, and no mobile holes in region *C*. As in the previous case, an internal electric field will be produced, which here acts from region *C*, (negatively ionized impurity atoms) towards region *B* (positively ionized impurity atoms). This field hinders the further diffusion of mobile holes from region *C* into region *B*, and of free electrons from region *B* into region *C*. We will indicate the junction formed in this way by B_2.

It is thus evident, that in the absence of an external voltage two junctions are formed in the transistor, each of which is the result of a transition from one type of germanium to the other.

17.3. The mode of operation of a *PNP* transistor

We will now explain the operation of a transistor on the basis of Fig. 109. The bias voltages applied to the transistor are obtained from two batteries, both of which can be switched off.

Switch 1 closed

The battery V_C produces an electric field in the transistor, which is directed from the negative pole of the battery through the transistor to the positive pole, and has a value E_C. This field will reinforce the effect of the internal electric field e_2 in the junction B_2 and as a result, the electrons from region *B* cannot diffuse into region *C*, nor can the mobile holes diffuse from region *C* into region *B*. Consequently, junction B_2 is non-conductive, i.e. the junction between the collector and base is blocked.

Switch 2 closed

Battery V_B produces an electric field E_B between regions *B* and *A* which is directed, in the external battery circuit, from the negative pole to the positive pole, thus from *B* to *A*. This external field E_B opposes the internal field e_1 in the junction B_1 and encourages diffusion, so that electrons can diffuse into region *A*, and mobile holes into region *B*. Junction B_1 is thus conductive, and a very large number of holes move into region *B*. As this region is very thin, and contains relatively few free electrons, there is little chance that these holes from region *A* will recombine with free electrons from region *B*.

Consequently there is now a large surplus of holes below junction B_2, and these will be able to diffuse from region *B* into region *C*. This will in fact happen as a result of the external field E_C, the direction of which is such that it moves the holes from region *B* into region *C*. In this way, junction B_2 also becomes conductive, and a current will flow through the V_C circuit. The size of this current depends on the hole density in regions *A* and *C*. Since the holes move via the base, a number of them will be filled up by free electrons in region *B*. An electron current will start to flow in the circuit which is supplied by V_B, so as to make up for the free electrons which have been lost in this way in the base (region *B*).

It is thus evident that no current can flow between the collector and the emitter of a transistor, unless a current is also flowing between the base and the emitter. We will now investigate the effect of the polarities of the voltages V_C and V_B.

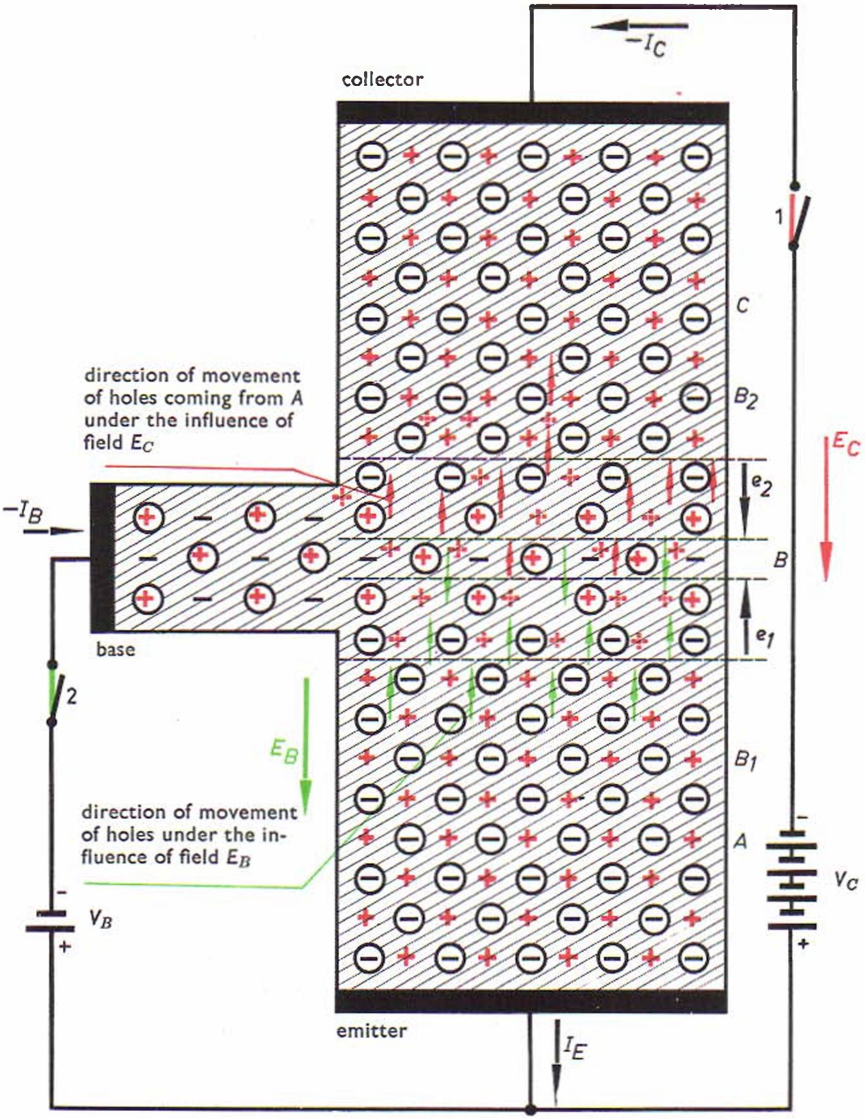
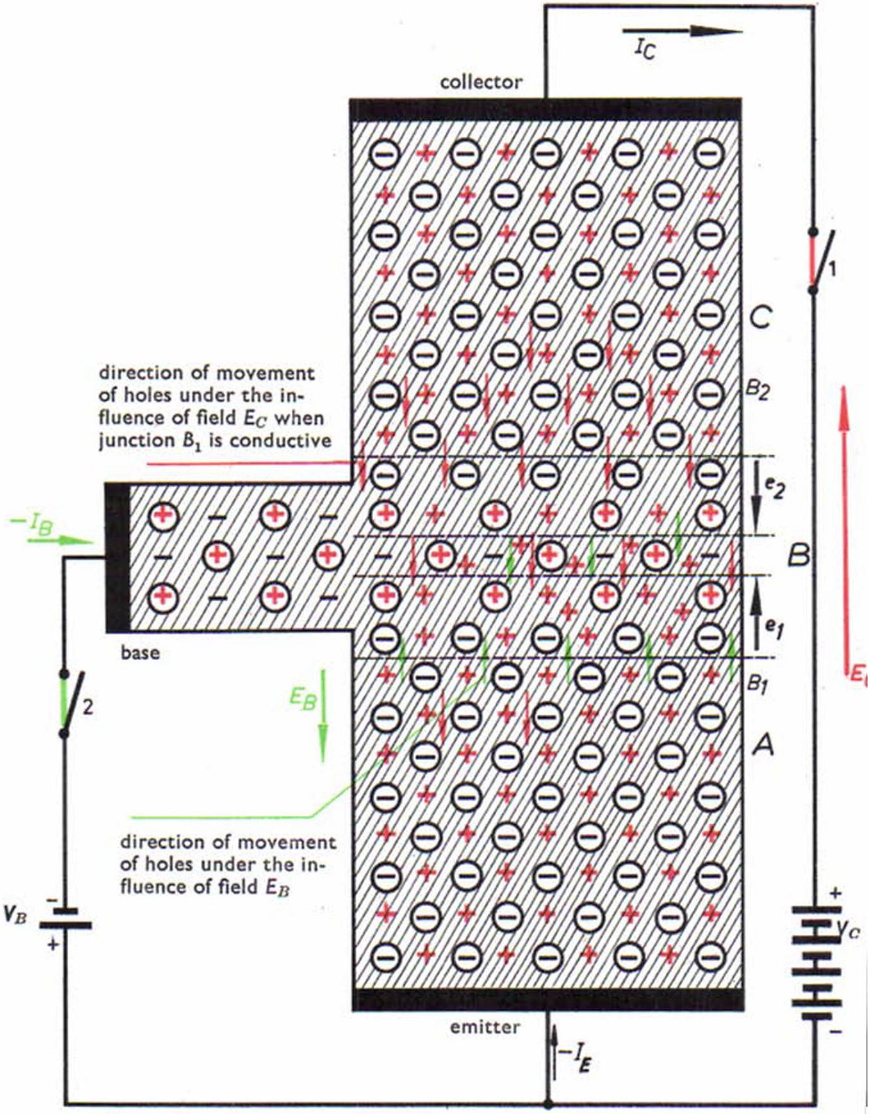

$-I_C$
collector
1
C
B_2
E_C
direction of movement of holes coming from A under the influence of field E_C
e_2
$-I_B$
B
e_1
base
2
E_B
B_1
direction of movement of holes under the influence of field E_B
A
V_C
V_B
emitter
I_E

Fig. 109


I_C
collector
1
C
B_2
e_2
B
e_1
B_1
A
E
$-I_B$
base
2
E_B
V_B
-
+
V_C
+
-
emitter
$-I_E$

direction of movement of holes under the influence of field E_C when junction B_1 is conductive

direction of movement of holes under the influence of field E_B

$V_B < V_C$

Fig. 110

17.4. Reversal of the voltage V_C

Let us assume that the transistor is connected as in Fig. 110. If switches 1 and 2 are open, there is no voltage between the collector and emitter and between the base and emitter of the transistor. The two junctions B_1 and B_2 prevent diffusion between regions A and B and regions B and C of the transistor.

Switch 1 closed

The battery V_C produces an electric field E_C in the transistor, which is directed from the negative pole to the positive pole of the battery. Like the internal electric field e_1 which acts across junction B_1, this field prevents diffusion between regions A and B. Junction B_1 is thus non-conductive and the transistor remains blocked.

Switch 2 closed

The battery V_B gives rise to an electric field E_B between regions B and A, which is directed from the negative pole of the battery to the positive pole, that is from B to A. This external field opposes the internal field e_1 across junction B_1. As the field E_B encourages diffusion, holes will be able to diffuse from region A into region B, and electrons from region B into region A. Junction B_1 is thus conductive.

Similarly, the external field E_C opposes the internal field e_2 in junction B_2, so that holes can also diffuse from region C into region B. As a result, a heavy current will flow through the base, and the diffusion of holes from B_1 to B_2 will not be retarded in any way, so that the transistor is immediately destroyed.

17.5. Reversal of the voltages V_C and V_B

We will investigate what happens when both V_C and V_B are reversed, with reference to Fig. 111. If both switches 1 and 2 are open, there is no voltage applied between the collector and emitter and between the base and emitter of the transistor, so that the two junctions B_1 and B_2 prevent diffusion between regions A and B and regions B and C.

Switch 1 closed

The battery V_C gives rise to an electric field E_C in the transistor, which is directed from the negative pole to the positive pole. Junction B_1 is still non-conductive, so that the transistor remains blocked.

Switch 2 closed

For the base-collector junction B_2 of the transistor, batteries V_C and V_B are in series, so that the base – collector voltage is equal to:

$$-V_{BC} = -V_C + V_B.$$

This voltage difference causes an external electric field E_1; outside the batteries, this field is directed from the negative pole to the positive pole, thus from the base to the collector.

Under the influence of this field, holes can diffuse from region C into region B, and electrons from region B into region C. The excess of holes in junction B_1, which have come from region C, diffuse into region A under the influence of the field E_C. Consequently a heavy current flows through the circuit, which may destroy the transistor.

Although the transistor is never subjected to this situation under normal conditions, it must be protected against destruction if this should occur, and measures must be taken to prevent B_2 from conducting. The external field E_1 must not be directed in such a way that it reinforces the internal field e_2. Either we must see that voltage V_B is at least equal to voltage V_C, or the current must be limited to a safe value by means of external resistances.

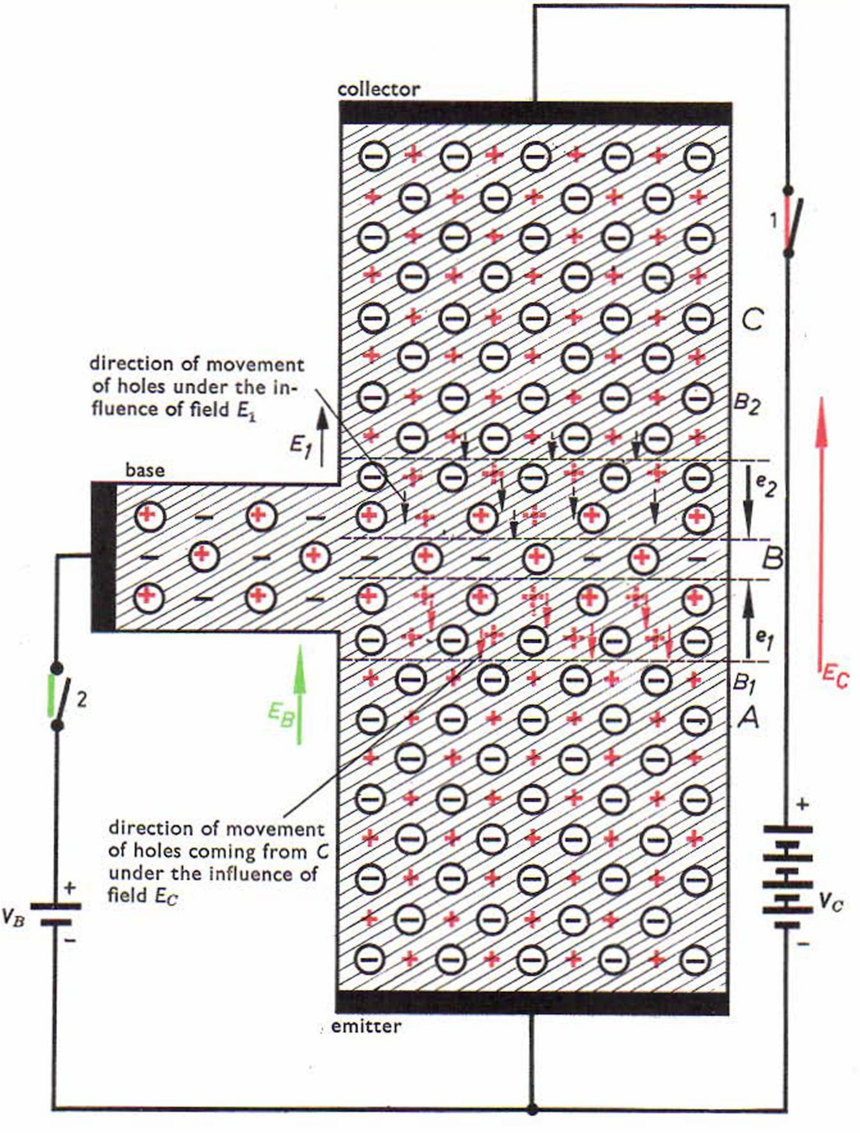

collector
1
C
direction of movement of holes under the influence of field E_1
E_1
B_2
base
e_2
B
e_1
E_C
B_1
2
E_B
A
direction of movement of holes coming from C under the influence of field E_C
+
V_B
-
+
V_C
-
emitter

Fig. 111

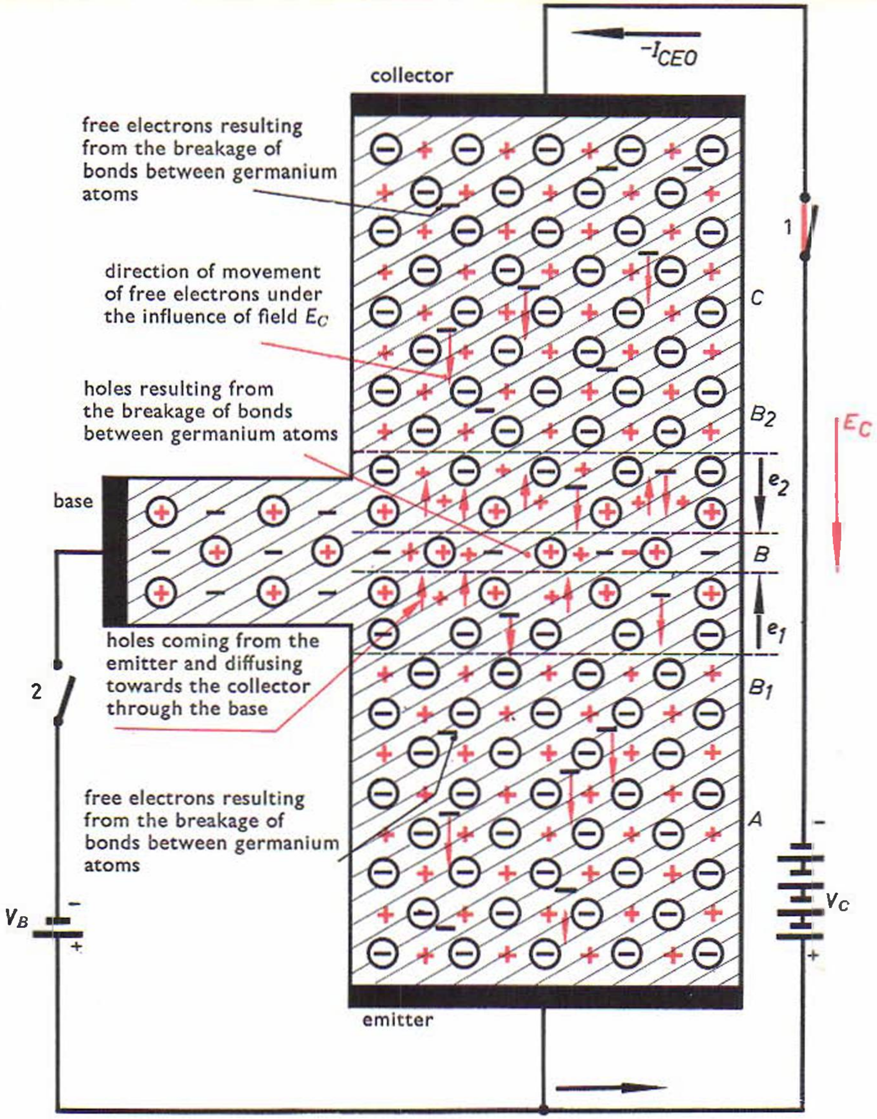

$-I_{CEO}$
collector
free electrons resulting from the breakage of bonds between germanium atoms
1
direction of movement of free electrons under the influence of field E_C
C
holes resulting from the breakage of bonds between germanium atoms
B_2
E_C
base
e_2
B
e_1
holes coming from the emitter and diffusing towards the collector through the base
2
B_1
free electrons resulting from the breakage of bonds between germanium atoms
A
V_B
V_C
emitter

Fig. 113

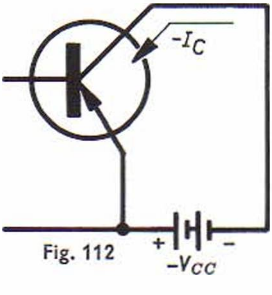


Fig. 112

17.6. Leakage current

We will explain the occurrence of leakage current on the basis of Fig. 113, which – as far as the circuit is concerned – corresponds to Fig. 112.

If switch 1 is closed the external field E_C will reinforce the internal field e_2 across junction B_2. It might be expected that under these conditions the transistor would not conduct a current, but a certain current does in fact flow in the circuit of battery V_C. This can be explained as follows.

In region *C*, which consists of *P* germanium, the impurity atoms will give rise to mobile holes. At room temperature, some bonds between germanium atoms will be broken, resulting in extra mobile holes and in particular, in extra free electrons. This is also true of region *A*.

In region *B*, which consists of *N*-type germanium, conduction is due to free electrons; these owe their existance to the presence of impurity atoms. At room temperature some bonds between germanium atoms are broken, giving rise to extra free electrons, and particularily to mobile holes. Under the influence of the field E_C, electrons can diffuse from region *C* into region *B* and mobile holes can diffuse from region *B* into region *C*.

The free electrons in the base which have come from the collector, diffuse under the influence of the field E_C into the emitter, so that B_1 is conductive. A still larger number of holes diffuse from the emitter to the base, and cross the base-collector junction, under the influence of field E_C. Consequently a relatively large current flows in the collector-emitter circuit of the transistor. This current depends on the number of broken bonds between germanium atoms; this number and therefore the leakage current also, will increase with increasing temperature.

CHAPTER 18

NPN transistors

Suppose that we have a block of *P*-type germanium with a sheet of *N*-type germanium alloyed to each side of it (Fig. 114). We will again term the three regions *A*, *B* and *C*. Regions *A* and *C* consist of *N*-type germanium and *B* consists of *P*-type germanium. The boundaries separating these three regions are indicated by the broken lines. In *NPN* transistors the percentage of impurity in the *N*-type region is greater than the percentage of impurity in the *P*-type region. The free-electron density in regions *A* and *C* is therefore much greater than the mobile-hole density in region *B*.

18.1. Formation of the *NP*-junction between regions *A* and *B*

The mobile holes in region *B* diffuse into region *A* where they will very quickly be filled up by the large surplus of free electrons which are present (Fig. 115). Similarly, free electrons will diffuse from region *A* into region *B*, where they will quickly fall into the large surplus of mobile holes. In the immediate neighbourhood of the junction there will no longer be any free electrons in region *A* or mobile holes in region *B*. Only positively ionized impurity atoms will remain in region *A* and negatively ionized impurity atoms in region *B*. In this way, an internal electric field e_1 is produced, directed from the *P*-type germanium towards the *N*-type germanium (thus from *B* to *A*). This field prevents the diffusion of free electrons from region *A* into region *B* and of mobile holes from region *B* into region *A*. In Fig. 115 this potential barrier is indicated by B_1.

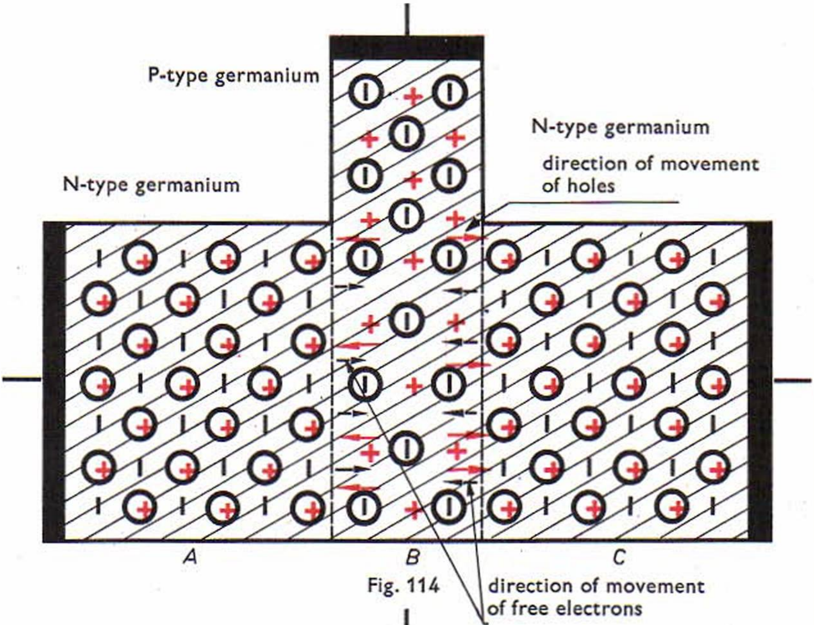

P-type germanium
N-type germanium
N-type germanium
direction of movement of holes
A
B
C
direction of movement of free electrons

Fig. 114

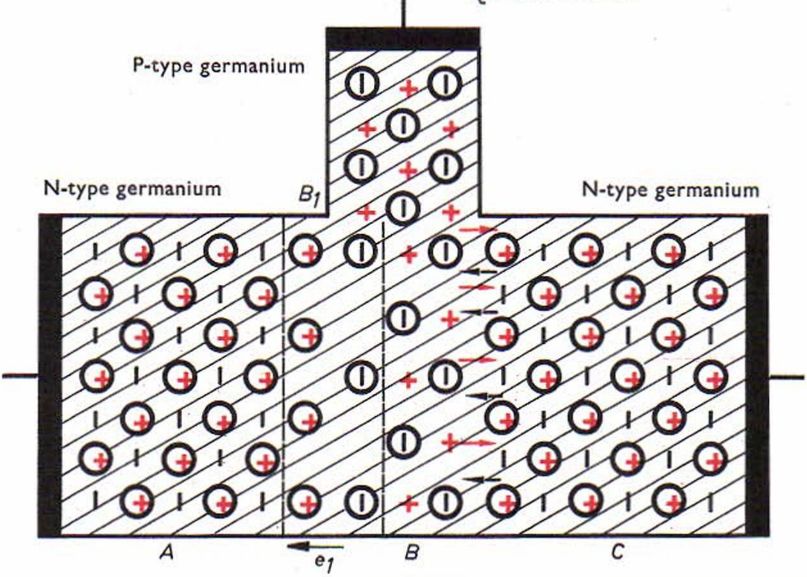

P-type germanium
N-type germanium
B_1
N-type germanium
A
e_1
B
C

Fig. 115

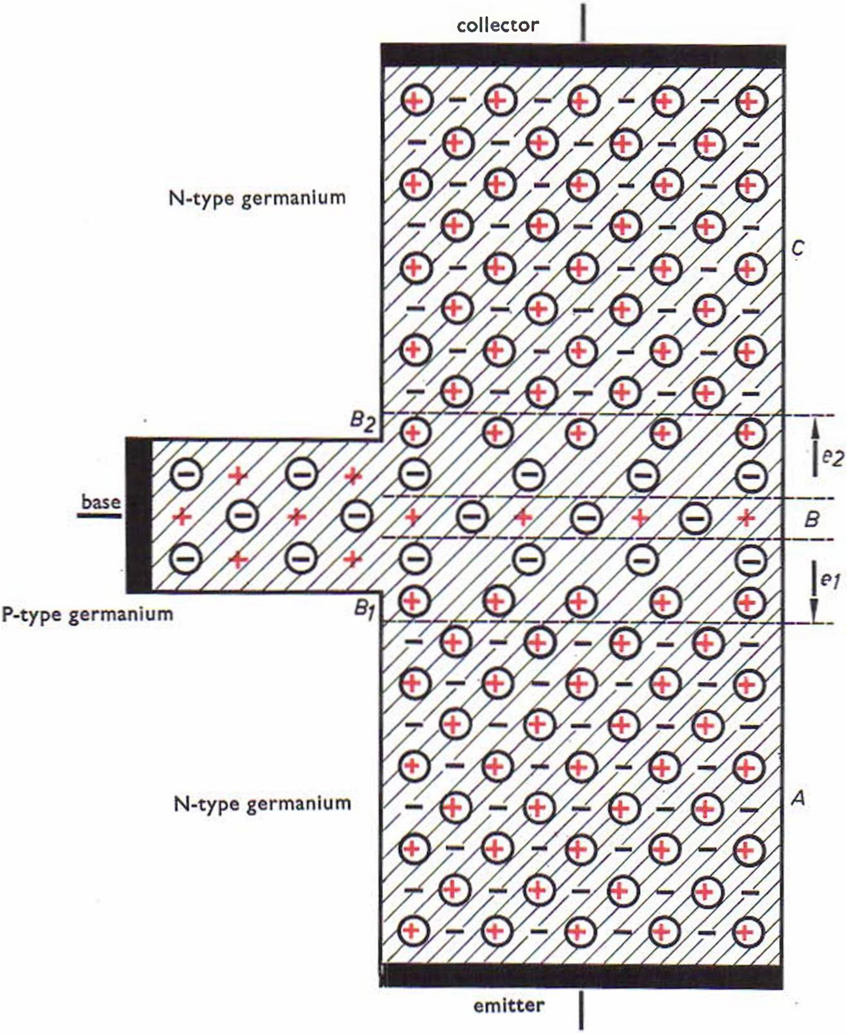

collector
N-type germanium
C
B2
e2
base
B
e1
P-type germanium
B1
N-type germanium
A
emitter

Fig. 116

18.2. Formation of the *PN*-junction between regions *B* and *C*

Region *B*, which consists of *P*-type germanium, is rich in mobile holes, while region *C*, which consists of *N*-type germanium, is very rich in free electrons (Fig. 116). The holes from region *B* diffuse into region *C* where they will very quickly be filled up by the free electrons, which are the majority carriers in this region. Similarily, the free electrons from region *C* will diffuse into region *B*, and will very quickly fall into mobile holes, which are very much in the majority there.

In the transition zone, there will no longer be any free electrons in region *C* or mobile holes in region *B*. The negatively ionized impurity atoms in region *B*, and the positively ionized atoms in region *C*, cause an internal electric field e_2 wich is directed from the *N*-type germanium towards the *P*-type germanium (i.e. from *C* to *B*). This field prevents the diffusion of free electrons into region *B* and of mobile holes into region *C*. We will indicate the potential barrier formed in this way by B_2.

In the absence of an external voltage, therefore, we see that two potential barriers are formed in the transistor, both being situated at the junction between two different types of germanium. It should be noted that these internal fields act in opposite directions.

18.3. The mode of operation of an *NPN* transistor

For the operation of an *NPN* transistor, see Fig. 117. The transistor is supplied by two batteries.

Switch 1 closed

Inside the transistor the battery V_C produces an electric field E_C, which acts from the negative pole towards the positive pole (from *A* to *C*). This field reinforces the internal field e_2 so that the transistor is non-conductive.

Switch 2 closed

The battery V_B produces an electric field E_B which is directed from the negative pole to the positive pole of the battery (from *A* to *B*); this field opposes the internal field e_1. Since E_B promotes diffusion, free electrons diffuse from region *A* into region *B* and mobile holes diffuse from region *B* into region *A*, so that the junction B_1 passes a current.

A surplus of free electrons builds up in region *B*. This region is only thin and contains relatively few mobile holes, so that the chance of recombination of electrons from region *A* with the holes in region *B* is fairly small. Consequently there are now a large number of free electrons below junction B_2. These electrons, which have come from region *A*, and which are very much in the majority in region *B*, will diffuse into region *C* under the influence of the external field E_C, so that the junction B_2 is also conductive. The current flowing through the circuit depends on the free-electron density in regions *A* and *C*.

If junction B_2 is conductive, some free electrons from region *A* fall into mobile holes in region *B*, while some holes from region *B* are filled up by electrons in region *A*. Consequently, a current will flow in the circuit which is supplied by V_B, in order to supplement the holes in region *B*. No current can flow between the collector and emitter of an *NPN* transistor unless there is a current flowing between the base and emitter of the transistor. If the polarity of the supply is reversed this may lead to destruction of the transistor if the collector is made negative with respect to the emitter and the base remains positive. If the polarities are reversed with respect to the emitter this will not damage the transistor, provided that V_B is lower than V_C.

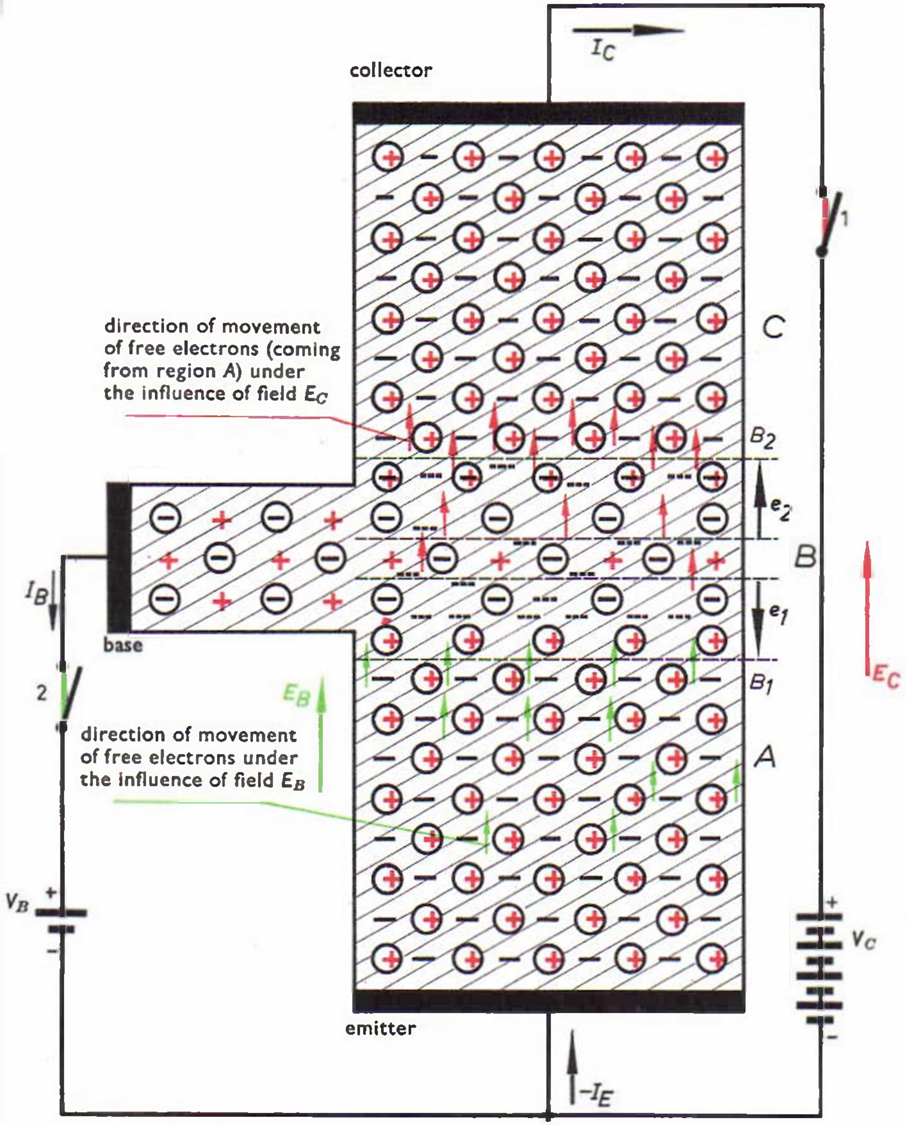

I_C
collector
1
C
direction of movement of free electrons (coming from region A) under the influence of field E_C
B_2
e_2
B
E_C
I_B
base
e_1
B_1
2
E_B
direction of movement of free electrons under the influence of field E_B
A
V_B
+
-
+
V_C
-
emitter
$-I_E$

Fig. 117

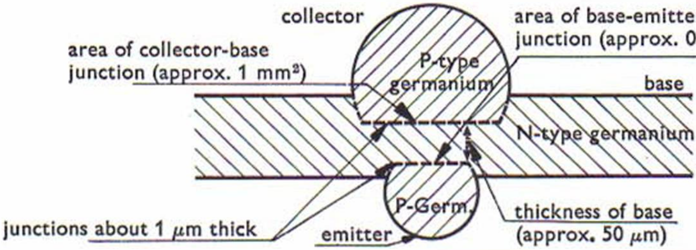

collector
area of collector-base junction (approx. 1 mm²)
P-type germanium
area of base-emitter junction (approx. 0.3 mm²)
base
N-type germanium
P-Germ.
junctions about 1 μm thick
emitter
thickness of base (approx. 50 μm)

Fig. 118

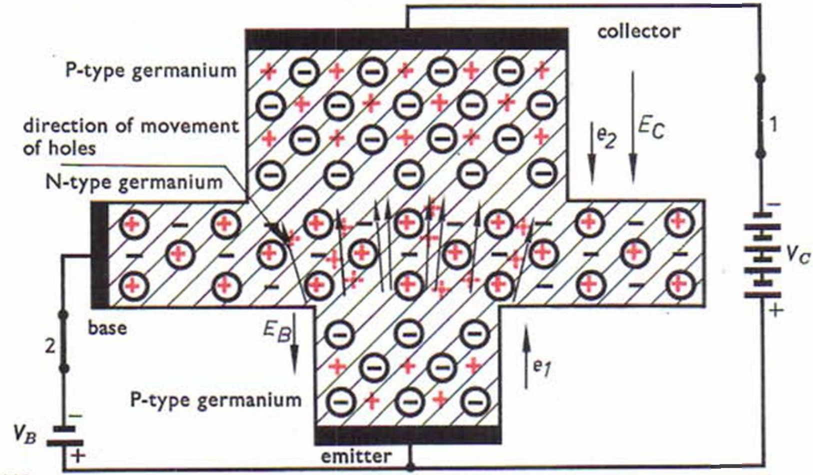

collector
P-type germanium
direction of movement of holes
N-type germanium
1
e2
EC
VC
base
2
EB
e1
P-type germanium
VB
emitter

Fig. 119

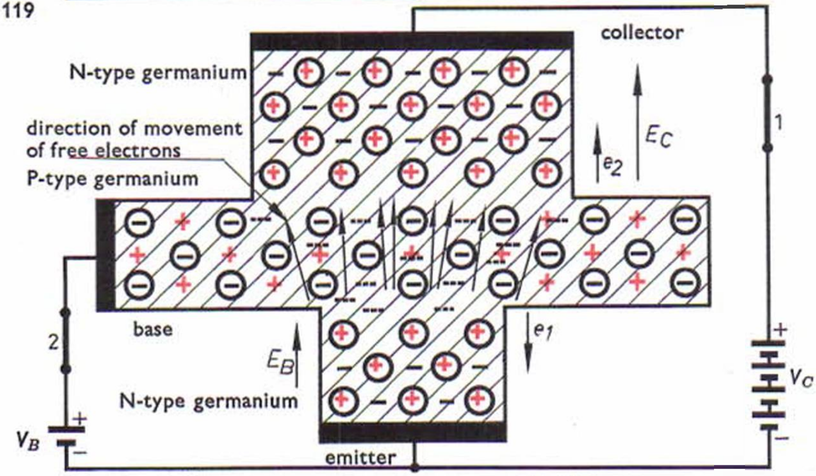

collector
N-type germanium
direction of movement of free electrons
P-type germanium
1
e2
EC
base
2
EB
e1
VC
N-type germanium
VB
emitter

Fig. 120

18.4. The construction of a junction transistor

Fig 118 represents a cross-section of a transistor. We see that the base is very thin (of the order of 50µm). The junctions are only about 1µm thick. It should also be noted that the area of the collector-base junction is much greater than that of the base-emitter junction (about three times as large).

In order to facilitate the above explanations no distinction was made in the corresponding figures between the dimensions of the two junctions. In fact, a *PNP* transistor should be represented as in Fig. 119. If switches 1 and 2 are closed the external field E_B will cause holes to diffuse from region *A* into region *B*, and electrons to diffuse from region *B* into region *A*. The holes coming from the emitter region diffuse within the base region in the directions indicated by the arrows in the schematic representation. To ensure that the collector region collects as many as possible of these holes, the area of the collector junction must be increased. This explains why the base-collector junction is always larger in area than the base-emitter junction.

The above example refers to a *PNP* transistor, but the same considerations are also valid for *NPN* transistors, as will be seen from Fig. 120,which represents a transistor of this type in approximately the correct proportions. The free electrons moving from the emitter region diffuse through the base region towards the collector region in the direction indicated by the arrows. In order to collect the largest possible number of electrons, the base-collector junction must thus be much greater in area than the base-emitter junction.

CHAPTER 19

Consideration of transistor characteristics

19.1. Commonly-used symbols

As transistors are usually used in the common-emitter configuration (see Fig. 121), we will limit ourselves to studying the characteristics referring to this circuit. We must be able to indicate the various currents and voltages in a transistor circuit by means of symbols, both for static conditions and for dynamic conditions.

1) Static conditions (direct voltages and currents)

The following symbols are used for static conditions:

$-V_{CE}$ for the collector-emitter voltage
$-V_{BE}$ for the base-emitter voltage
$-V_{CC}$ for the voltage of the battery which supplies the collector.
$-V_{BB}$ for the voltage of the battery which supplies the base.
$-I_C$ for the collector current.
$-I_B$ for the base current.
I_E for the emitter current.

Since the collector and the base are always at a negative voltage in relation to the emitter, their symbols are preceded by the negative sign.

In relation to the centre of the transistor, the base and collector currents are always in opposite directions referring to the emitter current (Fig. 122). The direction of the base and collector currents is also symbolized by the use of a negative sign.

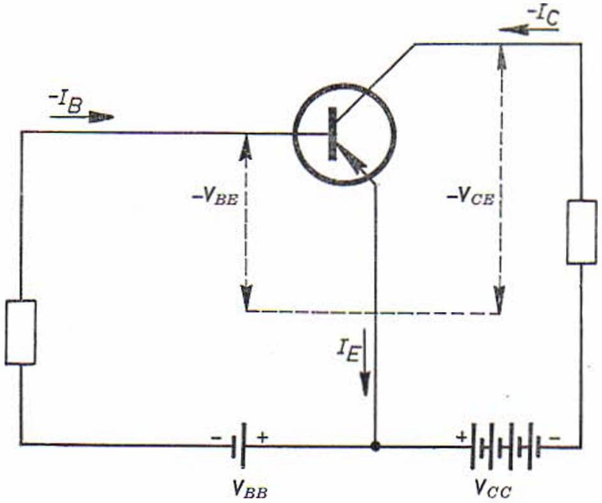

$-I_C$
$-I_B$
$-V_{BE}$
$-V_{CE}$
I_E
V_{BB}
V_{CC}

Fig. 121

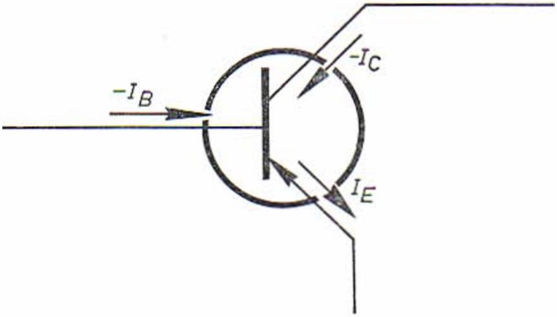


Fig. 122

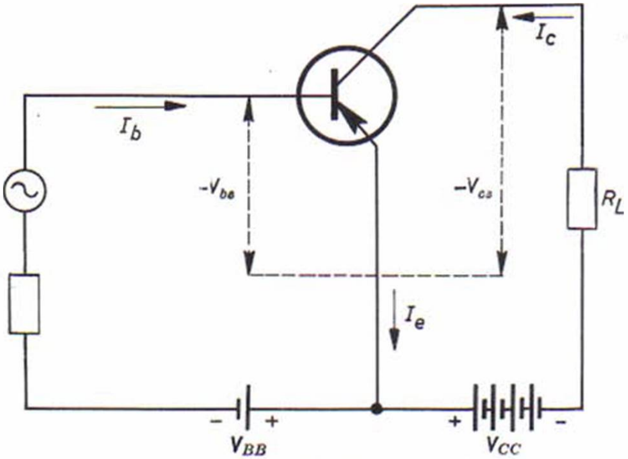

I_c
I_b
$-V_{be}$
$-V_{ce}$
R_L
I_e
− +
V_{BB}
+ −
V_{CC}

Fig. 123

2) Dynamic conditions (alternating voltages and currents)

The symbols normally used (Fig. 123) for dynamic conditions are as follows:

- v_{ce} for the collector – emitter voltage.
- v_{be} for the base – emitter voltage.
- i_c for the collector current.
- i_b for the base current.
- i_e for the emitter current.

The symbols defined in this way represent the r.m.s. values of the voltages and currents. The peak values of these quantities will be indicated as shown later. For example, the peak value of the collector-emitter voltage is indicated by the symbol V_{CM}.

19.2. Transistor characteristic curves

The object of this section is to explain how the principal characteristics of a transistor are obtained. Further on, we shall return to the way in which these characteristics can be used.

Transistor manufacturers provide a series of characteristics, some of which refer to the common-base configuration and some to the common-emitter configuration. There is a definite mutual relationship between the various voltages and currents associated with a transistor. The collector current is a function of the base current; the base current is determined by the base-emitter voltage, while under certain conditions, this voltage varies with the collector-emitter voltage. The characteristics which we shall now examine in more detail determine the mutual relationship which exists between these four parameters.

Fig. 124 shows the complete set of characteristics for a transistor in common emitter. We distinguish the following four quadrants:

top right $-I_C = f(-V_{CE})$,
top left $-I_C = f(-I_B)$,
bottom left $-I_B = f(-V_{BE})$,
bottom right $-V_{BE} = f(-V_{CE})$.

a) $-I_C = \mathrm{f}(V_{-CE})$ characteristic for various values of I_B

This family of curves (Fig. 125) represents the collector current $-I_C$ as a function of the collector-emitter voltage $-V_{CE}$, with the base current $-I_B$ as parameter. These characteristics can be determined by means of the circuit shown in Fig. 126. The collector-emitter voltage $-V_{CE}$ is read from the voltmeter connected between the collector and the emitter of the transistor. The collector current $-I_C$ is measured by means of a milliammeter included in the collector circuit. In the same way the base current is measured by means of a microammeter, included in the base circuit.

The collector-emitter voltage can be adjusted by means of the potentiometer R_1 and the base current by means of potentiometer R_2. The polarities of the supply voltages are indicated in Fig. 126. First of all we adjust $-I_B$ to 10μA by means of R_2; we then vary the collector-emitter voltage (from 0 to 10 V) by means of R_1, and note the values of $-I_C$ which correspond to various values of this voltage. If necessary we adjust $-I_B$ by means of R_2 so that this current is always 10μA.
Next we adjust $-I_B$ to 20μA for example, by means of R_2, and we again vary the collector-emitter voltage (from 0 to 10 V) by means of R_1. Once again we note the values of $-I_C$ for various values of $-V_{CE}$. These measurements can also be carried out for other values of $-I_B$, so that we finally obtain the family of characteristics shown in Fig. 125.

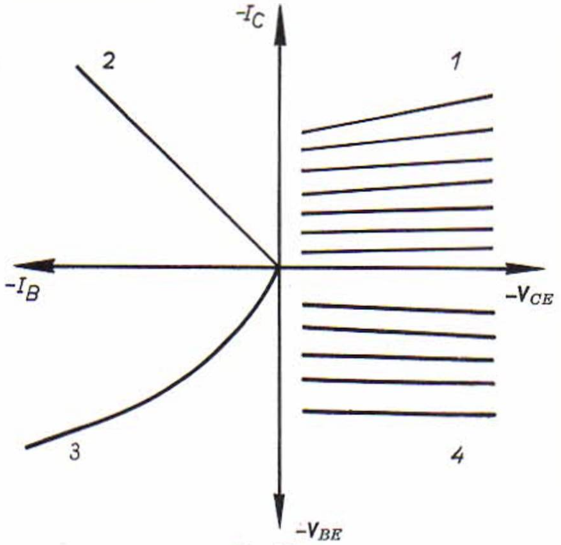


Fig. 124

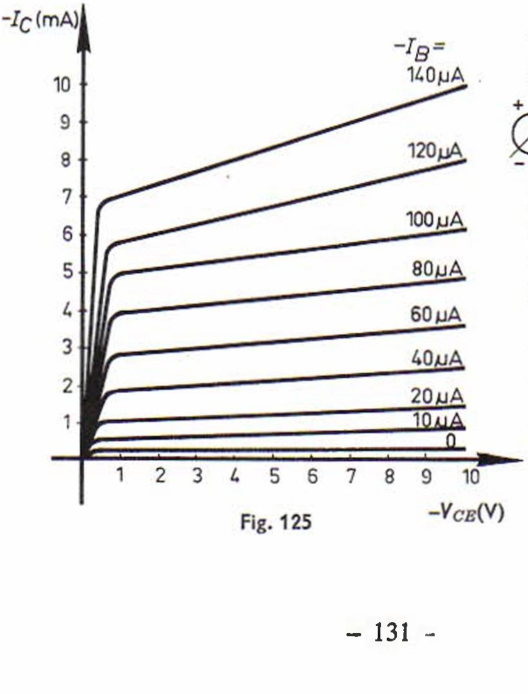
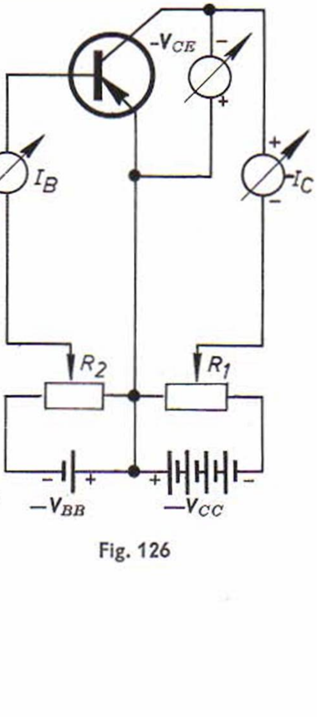


Fig. 125



$-V_{CE}$
I_B
I_C
R_2
R_1
$-V_{BB}$
$-V_{CC}$

Fig. 126

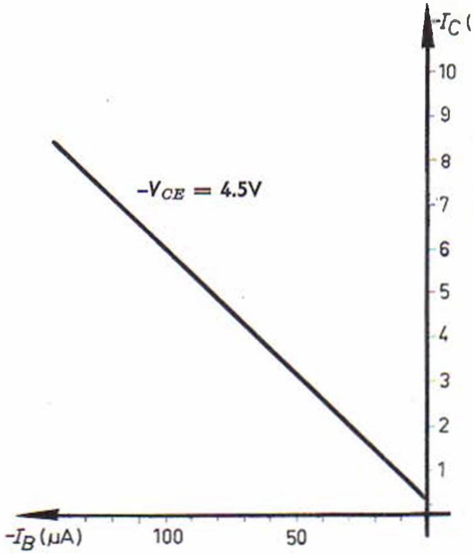


Fig. 127

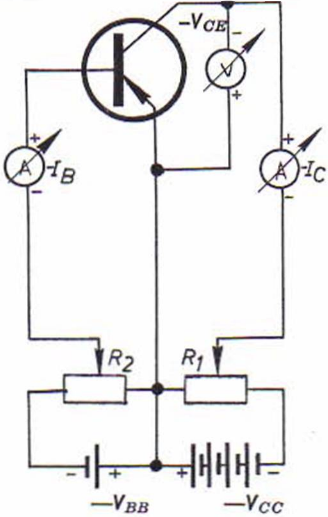

$-V_{CE}$
$-I_B$
$-I_C$
R_2
R_1
$-V_{BB}$
$-V_{CC}$

Fig. 128

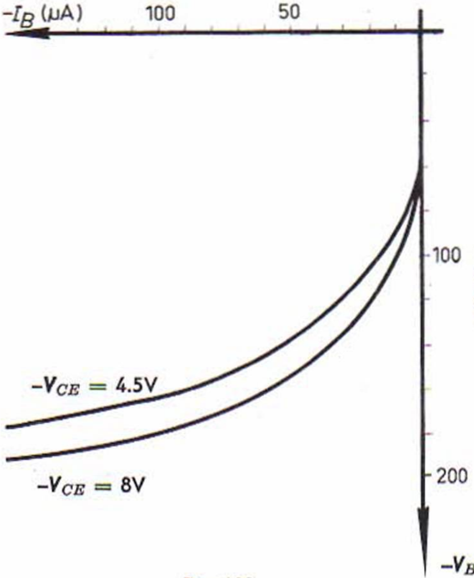


Fig. 129

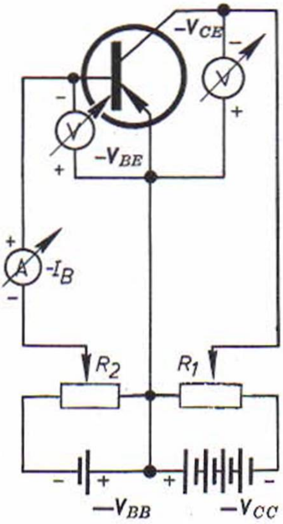

$-V_{CE}$
$-V_{BE}$
$-I_B$
R_2
R_1
$-V_{BB}$
$-V_{CC}$

Fig. 130

b) $-I_C = f(-I_B)$ characteristic

The curve $-I_C = f(-I_B)$ (see Fig. 127) represents the collector current as a function of the base current and if required can be recorded for various values of the collector-emitter voltage ($-V_{CE}$). To do this we use the circuit shown in Fig. 128.

The collector-emitter voltage ($-V_{CE}$) is measured by means of a voltmeter connected between the collector and the emitter of the transistor. The collector current is read from a milliammeter which is included in the collector circuit, and the base current is read from a microammeter included in the base circuit.

By means of R_1 we adjust $-V_{CE}$ to 4.5 V, for example, and then vary $-I_B$ by means of R_2, noting the values of $-I_C$ which correspond to various values of $-I_B$. The curve obtained in this way depends very little on the value of $-V_{CE}$, so that it is sufficient to determine the curve for a single value.

c) $-I_B = f(-V_{BE})$ characteristic for various values of $-V_{CE}$

The curve $-I_B = f(-V_{BE})$ (see Fig. 129) represents the base current as a function of the base-emitter voltage, with the collector voltage as parameter. This characteristic can be determined by means of the circuit shown in Fig. 130. The base-emitter voltage can be adjusted by means of R_2, and the collector-emitter voltage by means of R_1. The value of $-V_{BE}$ is measured by means of a millivoltmeter connected between base and emitter of the transistor.

We adjust $-V_{CE}$ to 4.5 V, for example, by means of R_1, and vary $-V_{BE}$ by means of R_2; the values of $-I_B$ corresponding to the various values of $-V_{BE}$ are then plotted. The second curve shown in Fig. 129 was obtained by repeating these measurements for $-V_{CE} = 8$ V.

d) $-V_{BE} = f(-V_{CE})$ characteristic for various values of $-I_B$

The curve $-V_{BE} = f(-V_{CE})$ (see Fig. 131) represents the base-emitter voltage as a function of the collector-emitter voltage, with the base current $-I_B$ as parameter. We determine this characteristic by means of the circuit shown in Fig. 132.

We adjust $-I_B$ to 10 µA, for example, by means of R_2, vary $-V_{CE}$ by means of R_1, and plot the values of $-V_{BE}$ corresponding to the different values of $-V_{CE}$. The family of curves in Fig. 131 was obtained by repeating the measurements for different values of the base current. It is customary to draw the axes of the characteristics (the abscissa and the ordinate) as shown in these figures (see Fig. 124).

19.3. Use of the characteristics

The $-I_C = f(-V_{CE})$ characteristic with $-I_B$ as parameter, and the $-I_B = f(-V_{BE})$ characteristic with $-V_{CE}$ as parameter are very important for studying the operation of a transistor. The first characteristic enables us to determine the behaviour of the output of the transistor and of its load circuit, while the second characteristic determines the behaviour of the input of the transistor and of the circuit that controls it.

The $-I_C = f(-I_B)$ characteristic and the $-V_{BE} = f(-V_{CE})$ characteristic with I_B as parameter are less important. The first enables us to investigate the effect of the input circuit on the output circuit, while the second determines the reaction of the output circuit onto the input circuit.

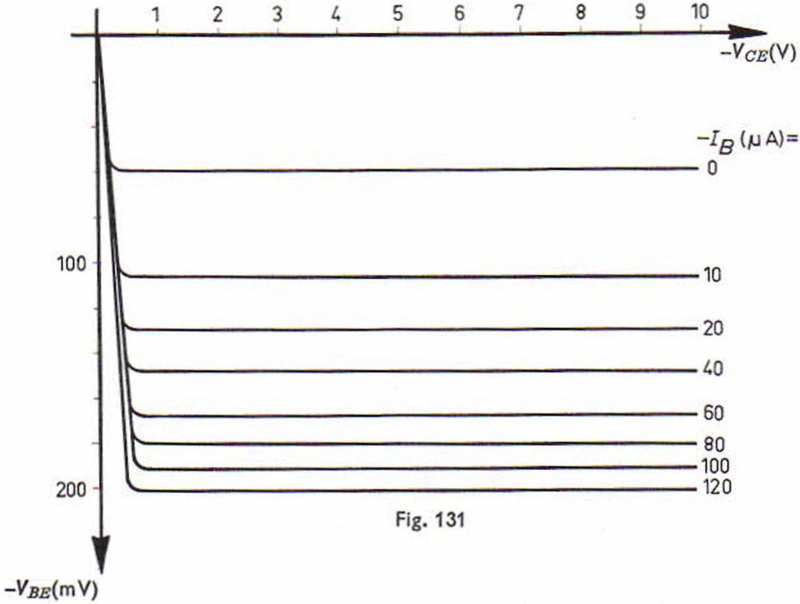


Fig. 131

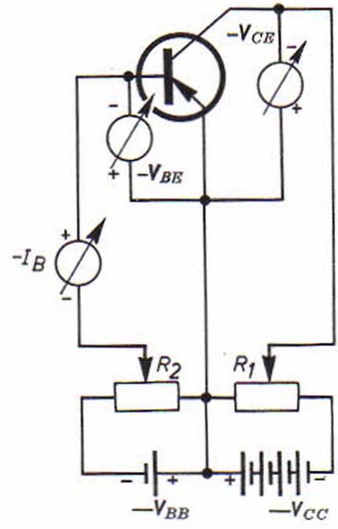

-V_CE
-
+
-
+
-V_BE
+
-I_B
-
R_2
R_1
- +
+ -
—V_BB
—V_CC

Fig. 132

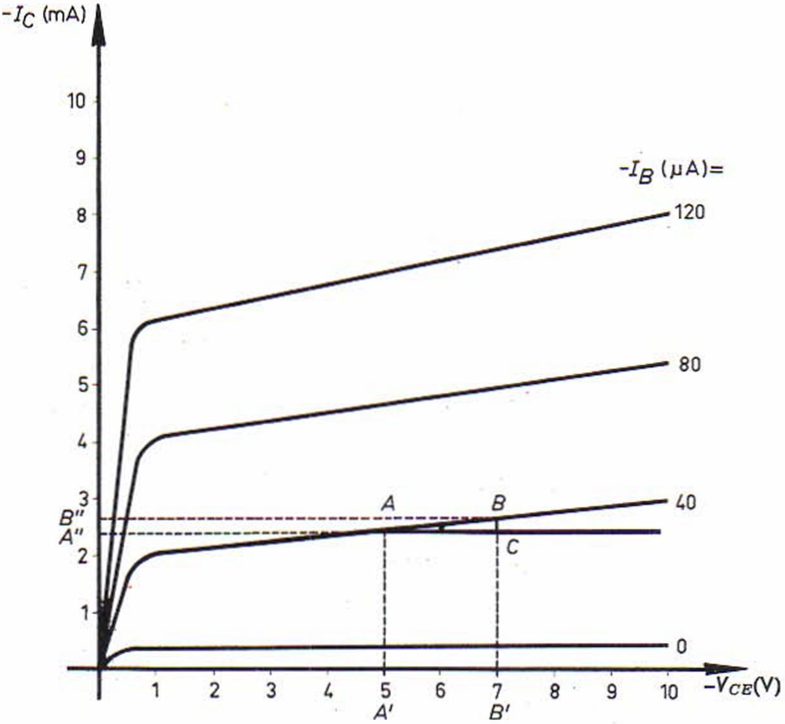


Fig. 133

a) $-I_C = f(-V_{CE})$ characteristic for various values of $-I_B$

With the aid of this characteristic it is possible to demonstrate the variation to which the following three important quantities are subject:

The output impedance of the transistor,
The current gain of the transistor,
The load line of the transistor.

Output impedance of the transistor

For our consideration of the output impedance of a transistor, we refer to Fig. 133. We will assume that $-V_{CE} = 5$ V; this voltage corresponds to point A' on the $-V_{CE}$ axis. At A' we erect the perpendicular to this axis and we will assume that the transistor works with a base current $-I_B$ of 40µA. The perpendicular cuts the curve for $-I_B = 40$µA at point A. This point is termed the working point of the transistor, and its projection on the $-I_C$ axis gives the point A''. If we produce the line $A''A$, this horizontal line forms an angle A with the curve.

Suppose that $-V_{CE}$ varies between 5 V and 7 V (point B'). The perpendicular to the $-V_{CE}$ axis at this point determines the point C on the horizontal line (extension of $A''A$) and the point B on the curve for $-I_B = 40$µA. If we project B on the $-I_C$ axis (point B''), we have:

$$\tan A = CB/CA — A''B''/A'B' = \Delta I_C/\Delta V_{CE} = 1/\mathrm{h}_{o(A)}$$

In which $\Delta I_C = 0.2$ mA $= 2 \times 10^{-4}$ A and $\Delta V_{CE} = 2$ V, while h_o represents the output impedance of the transistor, namely:

$$\mathrm{h}_{o(A)} = \Delta V_{CE}/\Delta I_C = 2/(2 \times 10^{-4}) = 10^4 = 10\ \mathrm{k}\Omega.$$

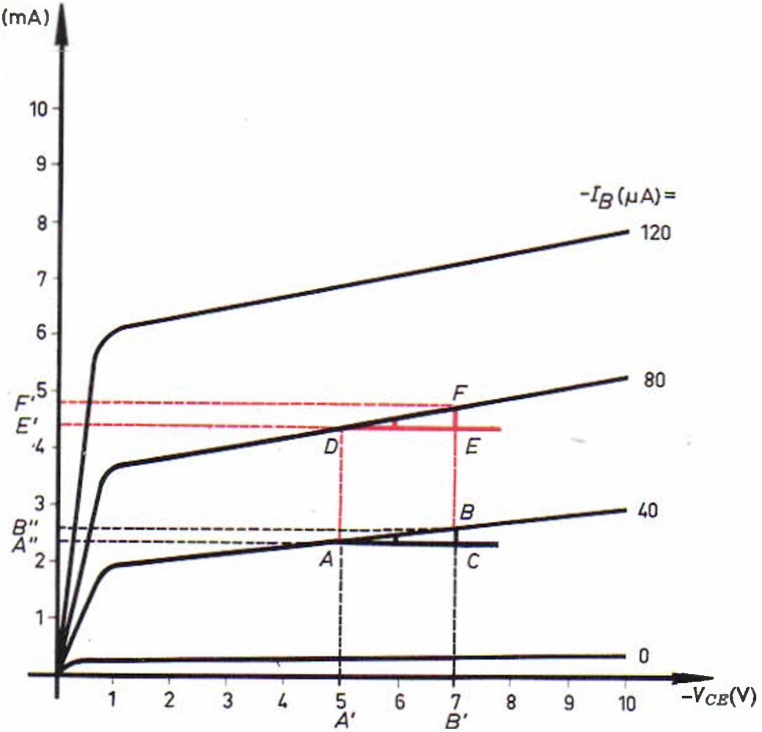
The perpendicular erected at point A' determines a point D on the $-I_B$ curve for 80µA (Fig. 134). The projection of this point D on the $-I_C$ axis gives the point E', and the perpendicular erected at point B' cuts the curve at point F, the projection of which on the $-I_C$ axis gives the point F'. The angle D between this curve and the horizontal is given by its tangent:

$$\tan D = EF/DE = E'F'/A'B' = \Delta I_C/\Delta V_{CE} = 1/\mathrm{h}_o,$$

where $A'B'$ represents the variation of the collector-emitter voltage (ΔV_{CE}), and $E'F'$ represents the corresponding variation (ΔI_C) of the collector current. In this case $\Delta V_{CE} = 2$ V and $\Delta I_{CE} = 0.3$ mA $= 3 \times 10^{-4}$ A, so that: $\mathrm{h}_{o(D)} = \Delta V_{CE}/\Delta I_c = 2/(3 \times 10^{-4}) = 6.7\ \mathrm{k}\Omega$.

In both cases ΔV_{CE} is the same, but ΔI_C is greater for angle D than for angle A, because $E'F'$ is greater than $B''A''$. As the tangent of angle D is greater than that of angle A, $1/\mathrm{h}_{o(D)}$ is greater than $1/\mathrm{h}_{o(A)}$. This shows that the output impedance of the transistor for $-I_B = 80$µA (at $-I_C = 4.5$ mA) is lower than for $-I_B = 40$µA (at $-I_C = 2.25$ mA).

For a given voltage, therefore, the output impedance of a transistor is determined by the angle between the characteristic and the horizontal, that is by the slope of the characteristic at the working point. Our consideration of these variations shows that the output impedance decreases as the collector current $-I_C$ increases. This follows from the fact that the angle between the characteristic and the horizontal increases with the value of $-I_C$ so that the slope also increases, and the output impedance decreases.



-I_C (mA)
10
9
8
7
6
5
4
3
2
1
F'
E'
B''
A''
-I_B (µA) =
120
80
40
0
F
D
E
B
A
C
1
2
3
4
5
6
7
8
9
10
A'
B'
-V_{CE}(V)

Fig. 134

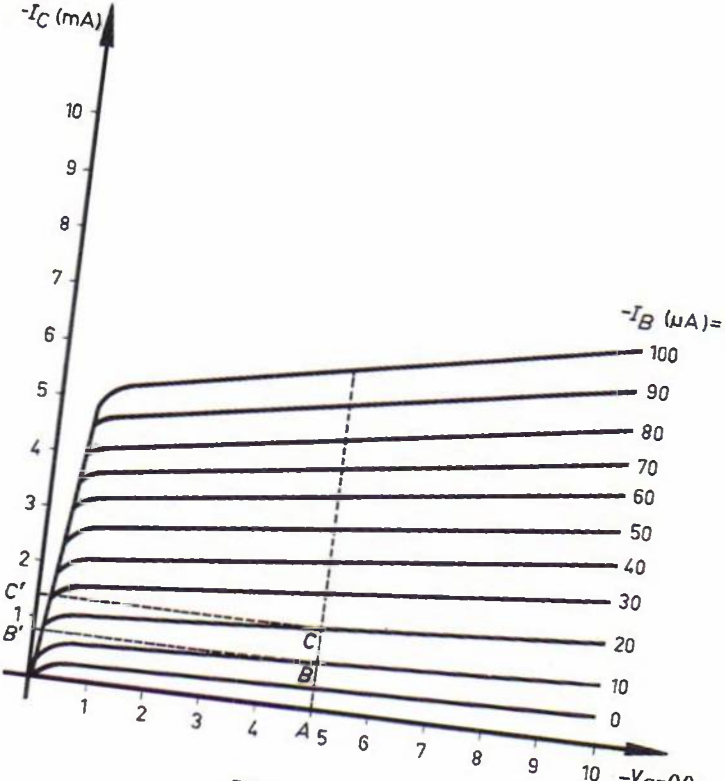

$-I_C$ (mA)
10
9
8
7
6
5
4
3
2
1
C′
B′
$-I_B$ (μA)=
100
90
80
70
60
50
40
30
20
10
0
C
B
A
1
2
3
4
5
6
7
8
9
10
$-V_{CE}$(V)

Fig. 135

Current gain of the transistor

The current gain of a transistor is defined as the ratio of the variation of collector current ΔI_C to the variation of base current ΔI_B which produces it. This definition applies to the transistor in common emitter. At the present time, this current amplification factor is represented by h_{fe} (previously by α' or α_{fe}, and in American publications by β):

$$\mathrm{h}_{fe} = \Delta I_C/\Delta I_B.$$

The characteristic from which this factor can be determined is shown in Fig. 135. Let us assume that $-V_{CE} = 5$ V. This voltage is represented by point A on the $-V_{CE}$ axis. If we erect the perpendicular at this point, this determines the point B on the curve for $-I_B = 10$ µA and point C on the curve for $I_B = 20$ µA.

Projection of points B and C on the $-I_C$ axis gives us the points B' and C' respectively. If the working point moves from B to C, the base current $-I_B$ increases from 10 µA to 20 µA and the collector current shifts from B' to C'. Now B' corresponds to $-I_C = 600$ µA and C' to $-I_C = 1200$ µA.

$B'C'$ represents the variation of collector current ,or in other words $\Delta I_C = 1200 — 600 = 600$ µA, so that:

$$\mathrm{h}_{fe} = B'C'/\Delta I_B = \Delta I_C/\Delta I_B = 600/10 = 60.$$

For power transistors, such as the OC 72 for example, the characteristics are of the form indicated in Fig. 136.

We will again assume that in the absence of a signal the collector-emittor voltage $-V_{CE} = 5$ V (point *A*). If we erect the perpendicular to the axis at point *A* it intersects the curve for $-I_B = 0.5$ mA at point *B* and the curve for $-I_B = 1$ mA at point *C*, the curve for $-I_B = 1.5$ mA at point *D*, and the curve for $-I_B = 2$ mA at point *E*.
If we now project points *B*, *C*, *D* and *E* on to the $-I_C$ axis we obtain the points *B'*, *C'*, *D'* and *E'*; *B'* corersponds to a collector current $-I_C$ of 40 mA, *C'* to a collector current $-I_C$ of 70 mA, *D'* to a collector current $-I_C$ of 90 mA and *E'* to a collector current $-I_C$ of 105 mA.

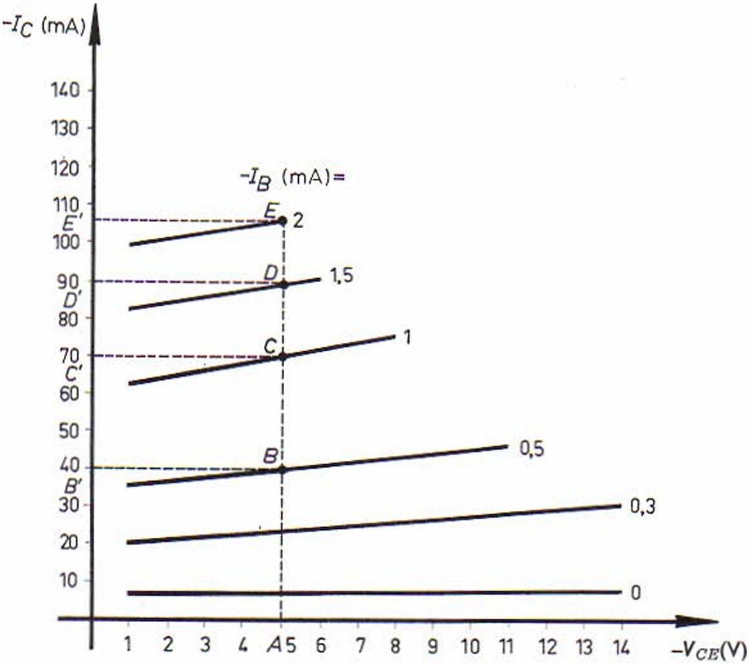
If the working point shifts from *B* to *C*, the base current increases from 0.5 mA to 1 mA, i.e. $I_B = 1 - 0.5 = 0.5$ mA; the collector current increases from 40 mA to 70 mA i.e. $\Delta I_C = 70 - 40 = 30$ mA. The current gain is thus equal to:

$$h_{fe} = \Delta I_C/\Delta I_B = 30/0.5 = 60.$$

If the working point shifts from *D* to *E*, the base current increases from 1.5 mA to 2 mA, i.e. $\Delta I_B = 1 - 0.5$ mA $= 0.5$ mA and the collector current then increases from 90 mA to 105 mA, i.e. $\Delta I_C = 105 - 90 = 15$ mA. In this case, the current gain equals:

$$h_{fe} = \Delta I_C/\Delta I_B = 15/0.5 = 30.$$

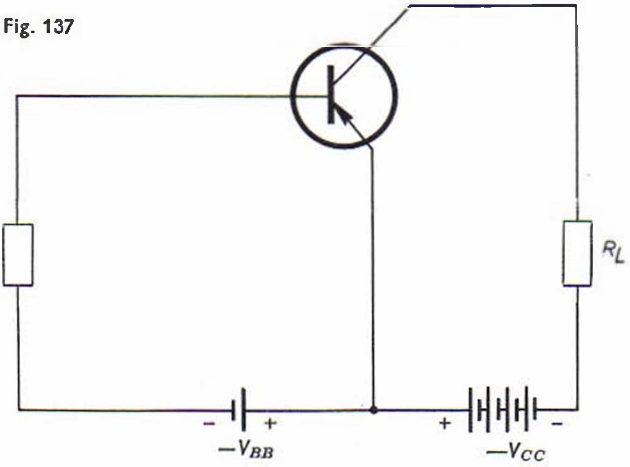
This shows that the current gain of a power transistor in grounded emitter decreases as the static collector current increases. This, of course, can also be seen from the $-I_C = f(-I_B)$ characteristic of this transistor.



$-I_C$ (mA)
140
130
120
110
E'
100
90
D'
80
70
C'
60
50
40
B'
30
20
10
$-I_B$ (mA)=
E
2
D
1,5
C
1
B
0,5
0,3
0
1
2
3
4
A
5
6
7
8
9
10
11
12
13
14
$-V_{CE}$(V)

Fig. 136

Fig. 137



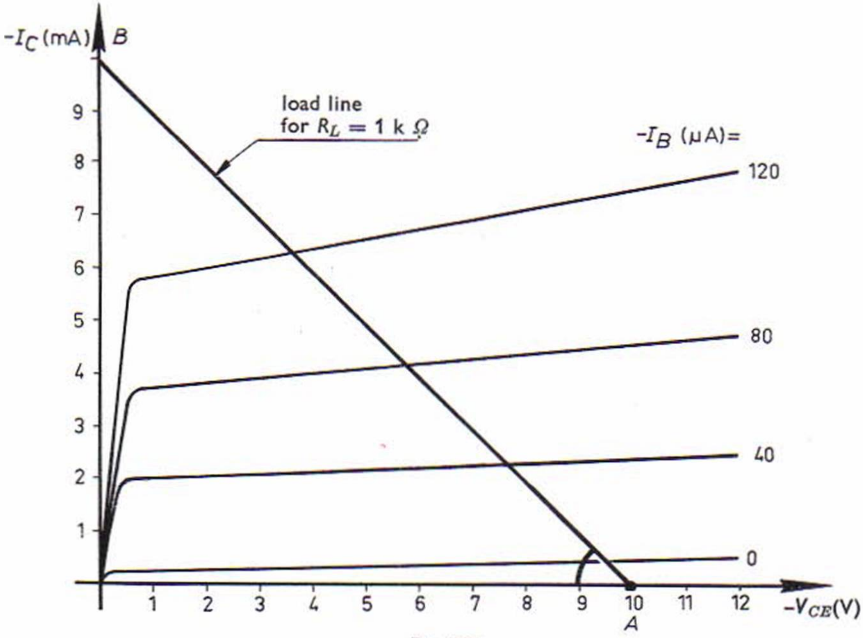


Fig. 138

Static load line

The static load line is a straight line which forms an angle A with the abscissa, such that $\tan A = 1/R_L$, in which R_L represents the load resistance of the transistor. We will explain this in more detail on the basis of Fig. 137, in which $-V_{CC} = 10$ V and $R_L = 1\text{k}\Omega$. We shall base our considerations on the family of characteristics shown in Fig. 138. The origin of the load line is at point A, which is given by:

$$-V_{CE} = -V_{CC} = 10 \text{ V}.$$

The load line AB and the abscissa $-V_{CE}$ form an angle A, whose tangent is equal to:

$$\tan A = \Delta I_C/\Delta V_{RL} = 1/R_L$$

In this equation V_{RL} must be expressed in volts, I_C in amperes and R_L in ohms. Consequently the tangent of angle A can be expressed directly as $1/R_L$, in which R_L is expressed in ohms, provided that the $-I_C$ scale is marked in amperes and the $-V_{CE}$ scale is marked in volts. However, if $-I_C$ is expressed in milliamperes R_L must be expressed in kilo-ohms. According to Fig. 138, $\tan A = 1/R_L = 1/1 = 1$, provided that 1 V on the abscissa corresponds to 1 mA on the ordinate.

The angle whose tangent is equal to 1 is 45°. In Fig. 138, therefore, the load line of the transistor, for a load resistance R_L of 1 kΩ, is represented by a straight line which forms an angle of 45° with the abscissa at point A.

We will now investigate the value of R_L on the position of the load line in the $-I_C = f(-V_{CE})$ characteristic. Suppose that $R_L = 0$. In that case:

$$\tan A = 1/R_L = 1/0 = \infty.$$

The angle whose tangent is infinite is 90°. In Fig. 140 therefore, this load line is represented by the perpendicular to the $-V_{CE}$ axis at point A (the red line). A shift of the working point along this straight line does indeed mean that the collector current varies without any change in the collector voltage:

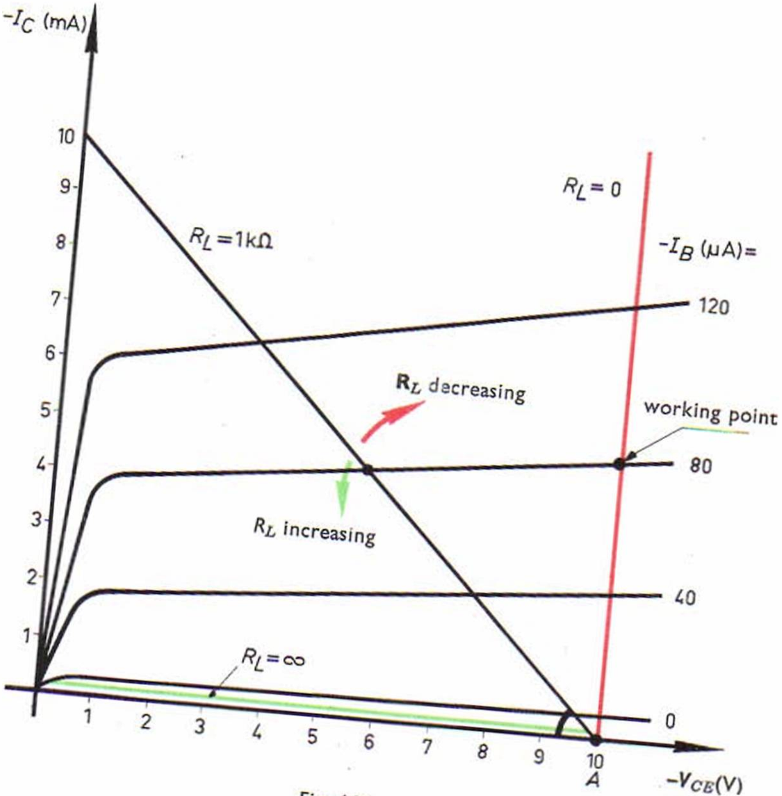
$$\Delta V_{RL} = R_L \,.\, \Delta I_C = 0.\Delta I_C = 0.$$

On the other hand if R_L is infinite,

$$\tan A = 1/R_L = 1/\infty = 0.$$

The angle whose tangent is 0, is 0°. The line which represents this angle (printed green in the figure) coincides with the abscissa and represents the load line of the transistor with an infinitely large load resistance R_L.

The figure shows that the slope of the load line decreases as R_L increases. The working point must lie on this load line; its position depends on a number of factors which we shall deal with later on.



$-I_C$ (mA)
10
9
8
7
6
5
4
3
2
1
$R_L = 1\text{k}\Omega$
$R_L = 0$
$-I_B$ (µA)=
120
80
40
0
R_L decreasing
R_L increasing
working point
$R_L = \infty$
1
2
3
4
5
6
7
8
9
10
A
$-V_{CE}$(V)

Fig. 140

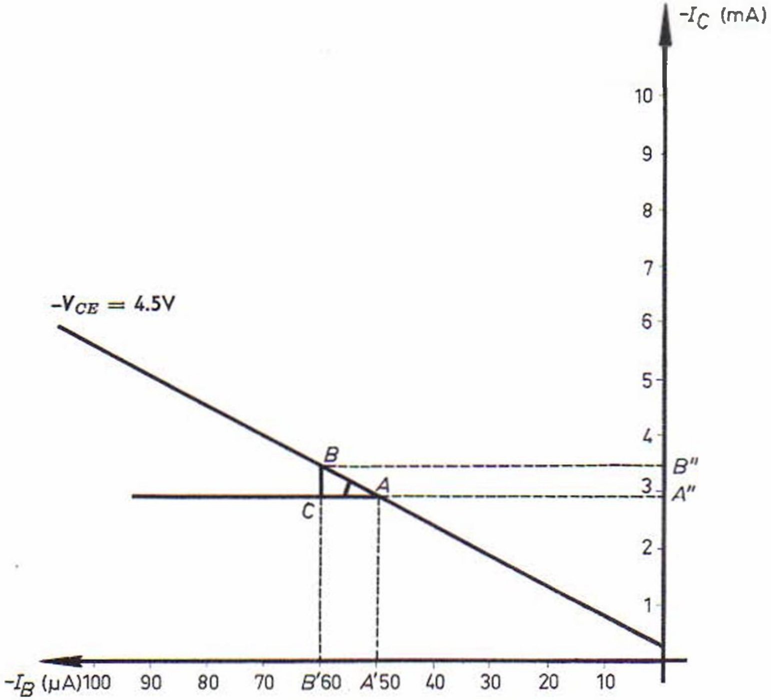

-I_C (mA)
10
9
8
7
6
5
4
3
2
1
-V_{CE} = 4.5V
B
A
C
B''
A''
-I_B (μA)100
90
80
70
B'60
A'50
40
30
20
10

Fig. 141

b) $-I_C = f(-I_B)$ characteristic

The relationship between the current gain and the operating conditions of a transistor can be investigated with the aid of the $-I_C = f(-I_B)$ characteristic.

Current gain of the transistor

We can determine the current gain of the transistor from Fig. 141. Suppose that the base current is 50 μA, corresponding to point A' on the $-I_B$ axis. At point A' we draw the perpendicular to the abscissa; this cuts the curve at point A. The projection of this point on the $-I_C$ axis, point A'', determines the collector current, which in this case is 2.8 mA.

Point B' on the abscissa corresponds to a base current of 60μA. The point of intersection of the perpendicular erected at this point with the curve determines point B. The projection of this point on the $-I_C$ axis gives point B'', corresponding to the collector current of 3.4. mA.

The current gain of a transistor in common emitter is given by the expression:

$$h_{fe} = \Delta I_C/\Delta I_B.$$

The extension of line $A''A$ cuts the line $B'B$ at C. The angle A between the characteristic and the horizontal is defined by its tangent, i.e.

$$\tan A = BC/CA = B''A''/B'A'.$$

in which $B'A'$ represents the change in the base current $\Delta I_B = 60 — 50 = 10$ μA, and $B''A''$ represents the change in the collector current $\Delta I_C = 3.4 — 2.8 = 0.6$ mA $= 600$ μA. Hence:

$$\tan A = h_{fe} = \Delta I_C/\Delta I_B = 600/10 = 60.$$

The slope of this characteristic is a measure of the current gain of the transistor.

When considering the relationship between the current gain and the operating conditions, we must distinguish between small-signal transistors and higher power transistors. For small-signal transistors the $-I_C = f(-I_B)$ characteristic is practically straight, whatever the value of $-V_{CE}$ (see Fig. 142). The slope of this characteristic is thus constant, which means that the current gain is also constant, independent of the value of the collector current.

If the transistor has to supply a greater power output, the characteristic is as shown in Fig. 143. Suppose that the base current $-I_B = 800$ µA, corresponding to point A' on the I_B axis. The perpendicular erected on the abscissa at this point cuts the characteristic at point A; the projection of point A on the $-I_C$ axis gives point A'' corresponding to $I_C = 52$ µA.

If $-I_B$ is 1 mA instead of 800 µA, (point B' on the $-I_B$ axis), the perpendicular erected at B' intersects the curve at point B; the projection of this point gives point B'' on the $-I_C$ axis, corresponding to $-I_C = 62$ mA. The extension of $A''A$ cuts the line $B'B$ at C. Now:

$$h_{fe} = \tan A = BC/CA = B''A''/B'A' = \Delta I_C/\Delta I_B = (62 - 52)/(1 - 0.8) = 50.$$

We will now consider the same variation for $-I_B = 2$ mA. (points D', D and D''), for which the corresponding collector current $-I_C = 100$ mA. If $-I_B$ increases to 2.2 mA (points E', E and E''), then the corresponding value of $-I_C = 104$ mA. In this case therefore, the current gain has dropped to:

$$h_{fe} = \tan A = EF/DF = D''E''/D'E' = 4/0.2 = 20.$$

The current gain of a power transistor thus decreases with increasing collector current.

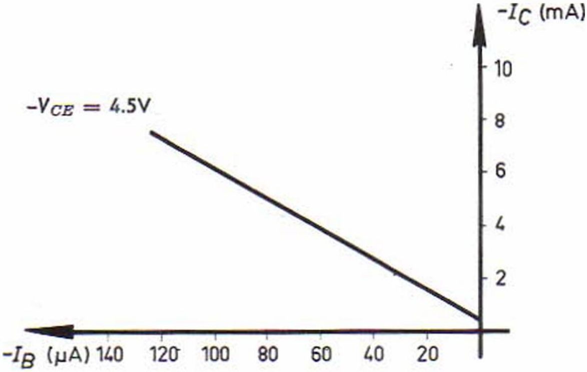


Fig. 142

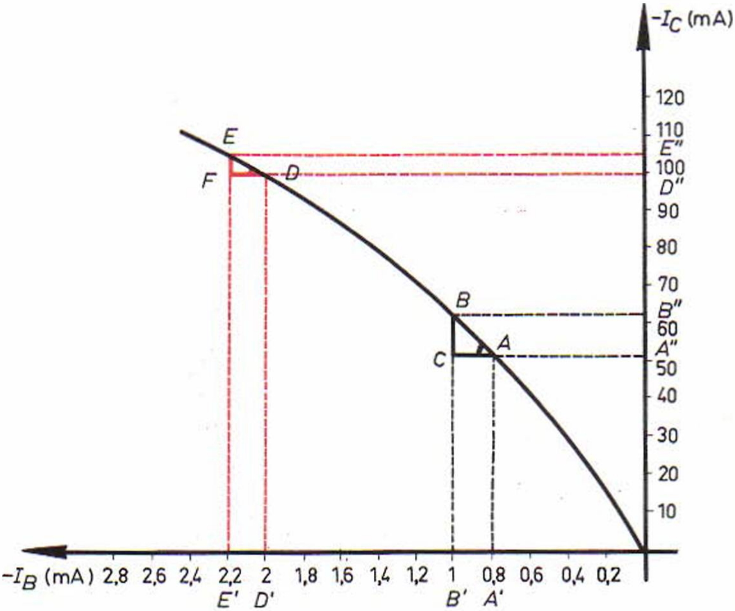


Fig. 143

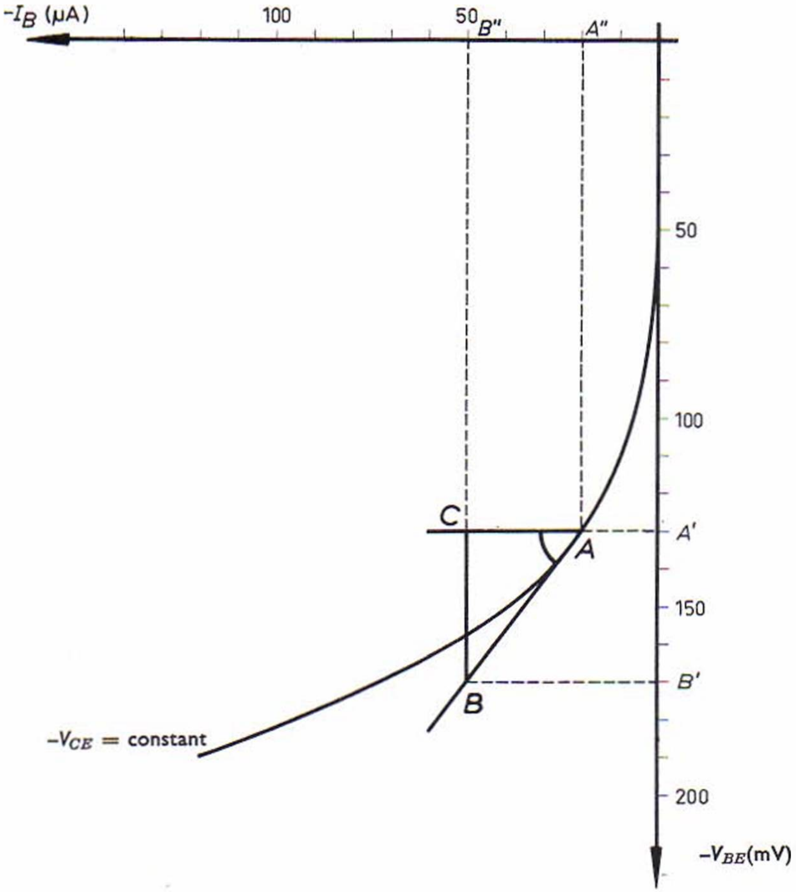


Fig. 144

c) $-I_B = f(-V_{BE})$ characteristic

The $-I_B = f(-V_{BE})$ characteristic enables us to study the variations of the following very important transistor quantities:

The input impedance of the transistor,
The input load line of the transistor.

Input impedance of the transistor

To study the input impedance h_i of a transistor, we start from Fig. 144. The transistor is biased so that the steady base-emitter voltage ($-V_{BE}$) is 130 mV: the base current $-I_B$ is then 20 μA. If we call the working point determined in this way A, and draw the tangent to the curve at this point and also a horizontal line through it, the angle A between the tangent and the horizontal is equal to the input impedance of the transistor for a given base – emitter voltage. The projections of point A on the ordinate and the abscissa, A' and A'', correspond to $-V_{BE} = 130$ mV and $-I_B = 20$ μA respectively.

Suppose that $-I_B$ increases from 20 μA to 50 μA (point B''). If we erect the perpendicular at point B'' this intersects the horizontal line through point A at C and the tangent to the characteristic at B. The projection of B on the $-V_{BE}$ axis gives point B'

$$\tan A = BC/CA = B'A'/A''B'' = \Delta V_{BE}/\Delta I_B.$$

The input resistance of the transistor equals the quotient of the variation of the base – emitter voltage, and the corresponding variation of the base current, i.e.:

$$h_{i(A)} = \Delta V_{BE}/\Delta I_B,$$

or in other words:

$$\tan A = \Delta V_{BE}/\Delta I_B = h_{i(A)}.$$

The slope of the $-I_B = f(-V_{BE})$ characteristic is thus a measure of the input impedance of the transistor for a given base-emitter voltage.

On the basis of Fig. 145 we will now investigate how the input impedance depends on the collector current. As in the previous case, we will again assume that the transistor is biased to give a base-emitter voltage of 130 mV (point A'). This voltage corresponds to a point A on the characteristic and the projection of point A on the $-I_B$ axis (point A'') gives a steady base current $-I_B$ of 20 µA. Through point A, we draw the tangent to the curve and a horizontal line.

If $-I_B$ increases from 20 µA to 50 µA (point B''), the new value corresponds to a point C on the horizontal line and a point B on the tangent to the characteristic. Point B is projected onto the $-V_{BE}$ axis, (point B'). From the characteristic of Fig. 145 we have:

$$\tan A = h_{i(A)} = BC/CA = A'B'/A''B'' = \Delta V_{BE}/\Delta I_B$$

$$= (17 — 13)10^{-2}/(5 — 2)\ 10^{-5} = 1300.$$

The input impedance of the transistor at a base-emitter voltage V_{BE} of 130 mV is thus 1300 Ω.

Now suppose that $-V_{BE} = 180$ mV (point D'), corresponding to a point D on the characteristic. The projection of this point on the $-I_B$ axis gives point D'', which shows that $-I_B$ now equals 70 µA. At point D we again draw the tangent to the curve, and a horizontal line (printed in red).

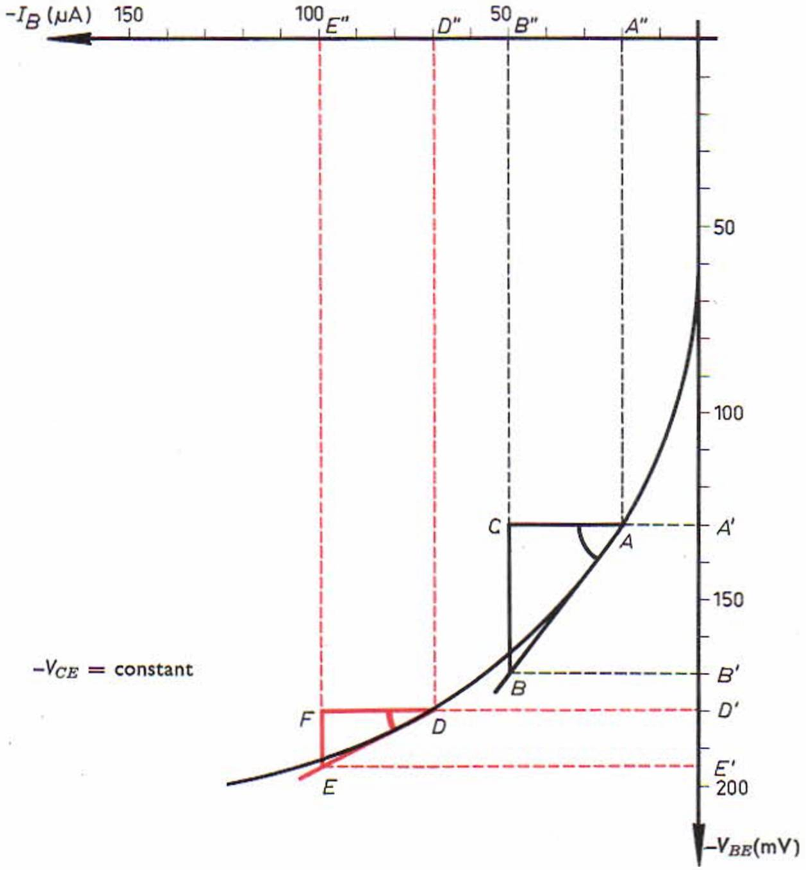
A point E'' on the $-I_B$ axis, corresponding to a base current of 100 µA, determines a point F on the horizontal line through point D and a point E on the tangent to the characteristic through point D. The projection of point E onto the $-V_{BE}$ axis gives point E', corresponding to 195 mV. This gives:

$$\tan D = h_{i(D)} = FE/FD = E'D'/E''D'' = \Delta V_{BE}/\Delta I_B$$

$$= (195 — 180)10^{-3}/(11 — 7)10^{-5} = 500\ \Omega.$$

It is thus evident that the input impedance h_i of the transistor decreases as the base current I_B increases. Since I_C increases with increasing I_B, it follows that as a general rule, the input impedance h_i decreases as I_C increases.

The sharp curvature of the input characteristic sets a limit to the linear operating range of a transistor. This curvature can be compared with that of the characteristic of a semiconductor diode in the forward direction.



$-I_B$ (μA)
150
100
E''
D''
50
B''
A''
50
100
A'
150
B'
D'
E'
200
$-V_{BE}$(mV)
C
A
B
F
D
E
$-V_{CE}$ = constant

Fig. 145

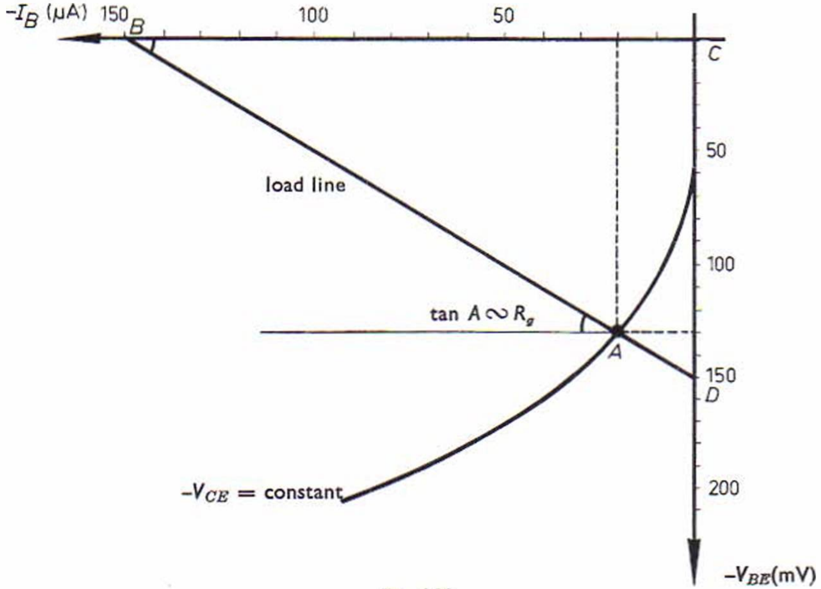
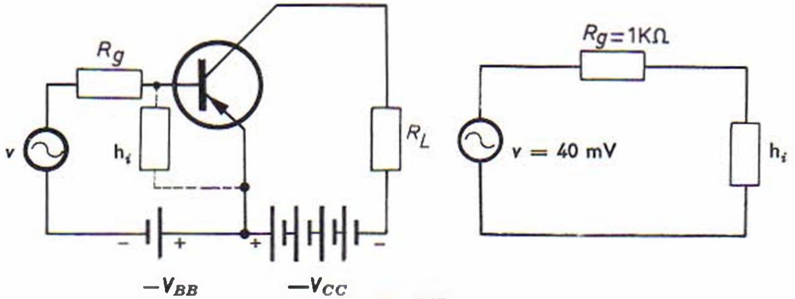


Fig. 146



R_g
v
h_i
R_L
–
+
+
–
$-V_{BB}$
$-V_{CC}$
R_g=1KΩ
v = 40 mV
h_i

Fig. 147

Input load line of the transistor

By the generator impedance R_g of a transistor circuit we understand the impedance or resistance of the generator which drives the transistor. We will examine this statement on the basis of Fig. 146. Suppose that A is the working point of the transistor, corresponding to a base-emitter voltage $-V_{BE}$ of 130 mV and a base current $-I_B$ of 20 μA. We will assume that the transistor is driven by a generator whose internal resistance is 1kΩ for example, and whose peak-to-peak no-load voltage is 40 mV. Fig. 147 represents the equivalent input circuit, in which a generator having no internal resistance supplies a voltage of 40 mV (peak-to-peak) to a circuit consisting of its internal resistance ($R_g = 1\text{k}\Omega$) in series with the input impedance of the transistor.

We will now draw the input load line in the $-I_B = f(-V_{BE})$ characteristic. To do this we draw a straight line through point A at an angle A to the $-I_B$ axis, such that $\tan A = R_g$ (In doing this, attention must be paid to the scales on the axis; if $-V_{BE}$ is expressed in volts and $-I_B$ in amperes, R_g is expressed in ohms. In the present case, $-V_{BE}$ is marked in millivolts and $-I_B$ is marked in microamperes, so that R_g is expressed in kilo-ohms).

If $\tan A = R_g = 1$, and the scales on the two axes are identical, $\angle A = 45°$. The input load line thus makes an angle of 45° with the horizontal through point A, and determines a point B on the $-I_B$ axis. As angle A is of course equal to angle B, we have:

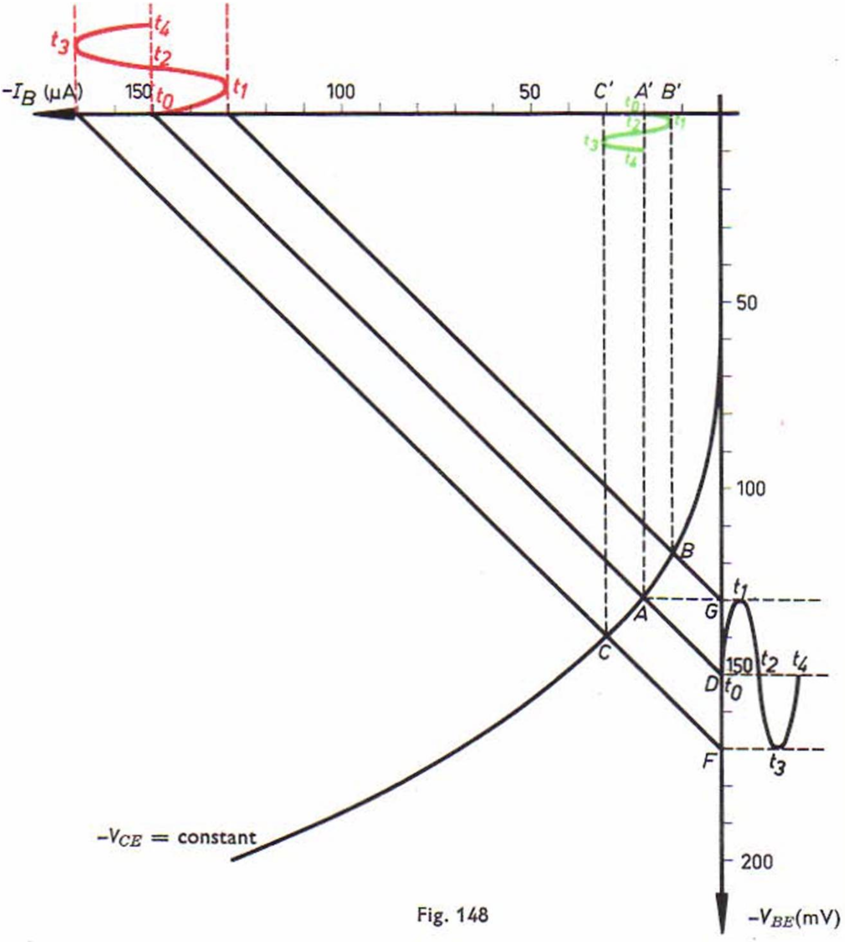
$$\tan B = CD/CB = \Delta V_{BE}/\Delta I_B = R_g.$$

If the generator impedance remains constant, variations of the bias voltage, and of the base current of the transistor, are reflected as a movement of the load line, parallel to itself.

The input load line which we have just drawn fixes a point D on the $-V_{BE}$ axis. We will now consider what happens when the generator voltage varies with respect to this point (Fig. 148).

From instant t_0 to instant t_1, this alternating voltage increases from zero to a maximum positive value; the base-emitter voltage then becomes less negative, and the load line is displaced parallel to itself, defining a point B on the characteristic. From instant t_1 to instant t_2 the alternating voltage decreases again to zero, and the load line returns to its original position (point A on the characteristic). From instant t_2 to instant t_3 the alternating voltage increases from zero to a maximum negative voltage; the base-emitter voltage becomes more negative and the load line is displaced parallel to itself, thus defining a point C on the characteristic. From instant t_3 to instant t_4 the alternating voltage decreases to zero once more and the base-emitter voltage returns to its original value, so that the load line takes up its original position again.

We thus see that the variations of the input voltage cause the input load line to move parallel to itself, thus defining three points on the characteristic. The projections A', B' and C' determine the variations of the base current of the transistor (printed in green). The red line in Fig. 148 represents the generator current which would flow if the generator was short-circuited. The slope of the input load line depends on the value of the generator impedance.



-I_B (µA)
150
100
50
C'
A'
B'
t3
t4
t2
t0
t1
t0
t2
t1
t3
t4
50
100
B
A
G
C
t1
150
t2
t4
D
t0
F
t3
-V_CE = constant
200
-V_BE(mV)

Fig. 148

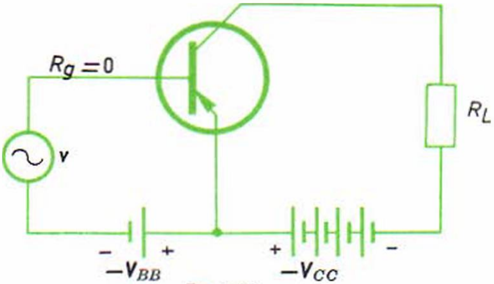

$R_g = 0$
R_L
v
− +
$-V_{BB}$
+ −
$-V_{CC}$

Fig. 149

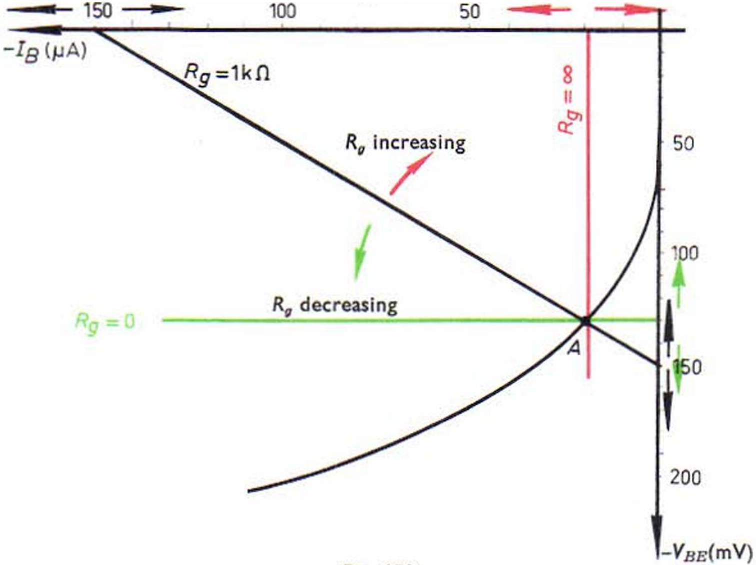


Fig. 150

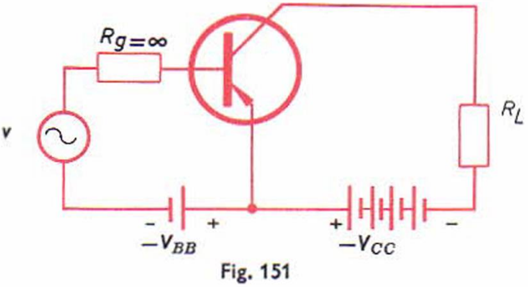

$R_g = \infty$
R_L
v
− +
$-V_{BB}$
+ −
$-V_{CC}$

Fig. 151

Let us first look at Fig. 149, in which the input load line $R_g = 0$ or, in other words, in which the generator which drives the stage has an infinitely small resistance (voltage source). In this case:

$$\tan A = R_g = 0.$$

The angle whose tangent is equal to 0 is 0°. This means that the input load line is horizontal through point A (printed green in Fig. 150).

We will now examine Fig. 151, in which the generator impedance $R_g = \infty$. This means that the transistor is driven by a generator whose internal resistance is infinitely large. In this case:

$$\tan A = R_g = \infty.$$

The angle whose tangent is infinitely large is 90°, and the input load line is now obtained by erecting the perpendicular to the horizontal line at point A (printed red in Fig. 150).

It is thus evident that the slope of the input load line decreases with decreasing generator impedance (lower internal resistance of the generator) but increases as this impedance increases (higher internal resistance of the generator). We can also draw the following conclusions from the curve in Fig. 150:

1) If the input load line approximates to the horizontal, the transistor is driven by voltage variations.
2) If the input load line approximates to the vertical, the transistor is driven by current variations.

d) $-V_{BE} = f(-V_{CE})$ characteristic

The degree of internal feedback of the transistor can be determined from the $-V_{BE} = f(-V_{CE})$ characteristic.

Internal feedback

The internal feedback of the transistor is given by the ratio of the variations of the base-emitter voltage to those of the collector-emitter voltage, i.e. the quotient $\Delta V_{BE}/\Delta V_{CE}$. We will explain this on the basis of Fig. 152. Suppose that $-V_{CE} = 5$ V (point A''). The perpendicular erected at this point determines point A on the curve $-I_B = 0$. The perpendicular dropped from this point to the $-V_{BE}$ axis gives point A'. Suppose that $-V_{CE}$ increases from 5 V to 10 V (point B''); the perpendicular erected at this point determines point C on the horizontal line through point A and B on the curve for $-I_B = 0$. The projection of the last point on the $-V_{BE}$ axis gives point B'. The angle A is determined by its tangent:

$$\tan A = CB/CA = A'B'/A''B'',$$

where $A'B'$ represents the variations ΔV_{BE} of the base-emitter voltage, and $A''B''$ represents that of the collector-emitter voltage ΔV_{CE}. It follows from this that the internal feedback of the transistor is:

$$\tan A = A'B'/A''B'' = \Delta V_{BE}/\Delta V_{CE}$$

This increases as the slope of the $-V_{BE} = f(-V_{CE})$ characteristic of the transistor increases.

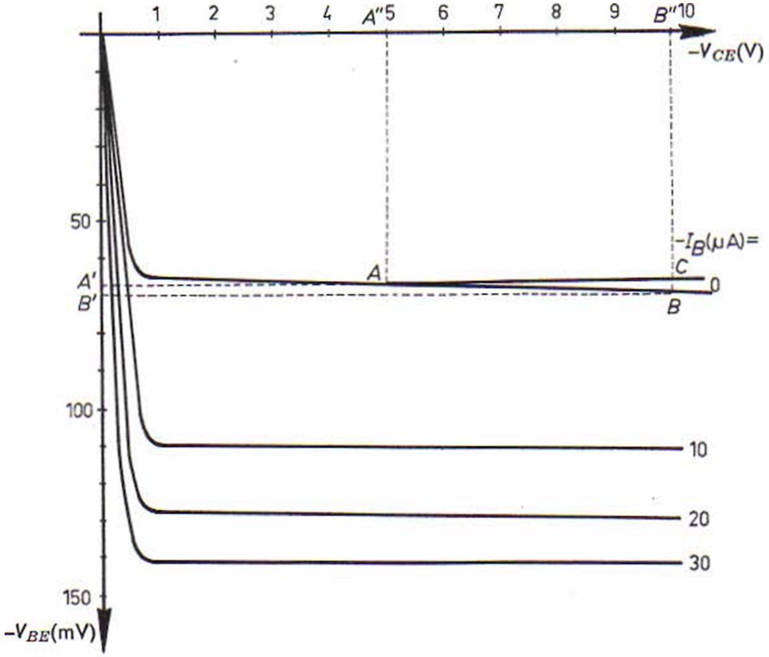


Fig. 152

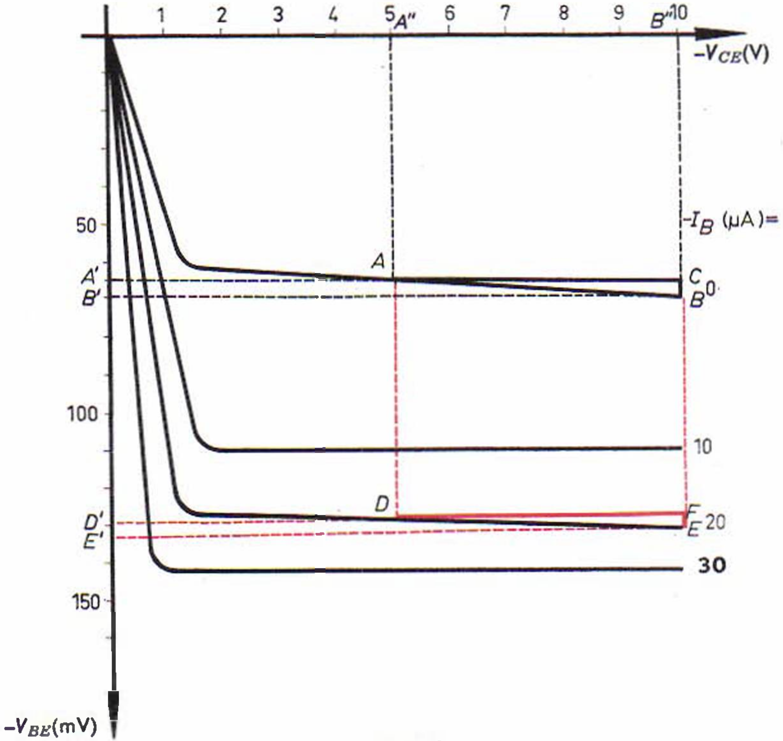

1
2
3
4
5
A''
6
7
8
9
B''
10
$-V_{CE}$(V)
50
100
150
$-V_{BE}$(mV)
$-I_B$ (μA)=
A
A'
B'
C
B
0
10
D
D'
E'
F
E
20
30

Fig. 153

Variation of the internal feedback

In Fig. 153 point A'' corresponds to a collector-emitter voltage $-V_{CE}$ of 5 V. The projection of this point on the curve for $-I_B = 0$ determines point A, whose projection A' on the $-V_{BE}$ axis corresponds to 65 mV. If $-V_{CE}$ increases from 5 V to 10 V (point B'' on the $-V_{CE}$ axis), the increased value corresponds to point C on the horizontal line through point A, and point B on the curve for $-I_B = 0$. The projection of B on the $-V_{BE}$ axis (B') gives a base-emitter voltage of 70 mV. As we have explained, the internal feedback is given by the tangent of angle A:

$$\tan A = CB/CA = A'B'/A''B'' = \Delta V_{BE}/\Delta V_{CE}$$

or, since $\Delta V_{BE} = 70 — 65 = 2$ mV $= 2 \times 10^{-3} V$ and $\Delta V_{CE} = 10 — 5 = 5$ V,

$$\tan A = \Delta V_{BE}/\Delta V_{CE} = 5 \times 10^{-3}/5 = 10^{-3}.$$

Point A'' corresponds to a point D on the curve for $-I_B = 20$ μA; the projection D' of D on the $-V_{BE}$ axis gives $-V_{BE} = 129$ mV. The projection of B'' on this curve (point E) corresponds to point E' on the $-V_{BE}$ axis, i.e. 133 mV. In this case, the internal feedback is:

$$\tan D = EF/DF = E'D'/A''B'' = \Delta V_{BE}/\Delta V_{CE},$$

or since we now have $\Delta V_{BE} = 133 — 129 = 4$ mV $= 4 \times 10^{-3}$ V, while ΔV_{CE} remains equal to 5 V;

$$\tan D = 4 \times 10^{3-}/5 = 8 \times 10^{-4}.$$

We may thus conclude that the internal feedback of a transistor decreases as the collector current increases.

e) $P_{C\,max} = f(T_{amb})$ characteristic

In designing a transistor amplifier one must take into account not only the four characteristics that we have just discussed, but also the characteristic which indicates the maximum permissible collector dissipation $P_{C\max}$ as a function of the ambient temperature T_{amb}. This characteristic is shown in Fig. 154.

Explanation of the characteristic

Suppose that the ambient temperature is 35° C. We erect a perpendicular at a point on the abscissa which corresponds to 35° C; this determines point A on the characteristic. The projection of this point on the ordinate (point A') gives the maximum permissible collector dissipation as 100 mW. An ambient temperature of 45° C corresponds to point B on the characteristic and point B' on the ordinate, which represents a maximum permissible collector dissipation of 75 mW. This means that the maximum permissible collector dissipation at 45° C is 75 mW. In the same way, we can read from the graph that the maximum permissible collector dissipation at a temperature of 65° C is only 25 mW.

It is thus evident that the maximum power which can be dissipated by a transistor depends on the ambient temperature, and decreases as this temperature increases. Now the maximum permissible collector dissipation sets a limit to the maximum output power of the transistor, which is therefore also temperature-dependent. This will be explained in more detail in the following discussion of the $P_{C\max} = f(T_{\mathrm{amb}})$ characteristic, in combination with the $-I_C = f(-V_{CE})$ characteristic.

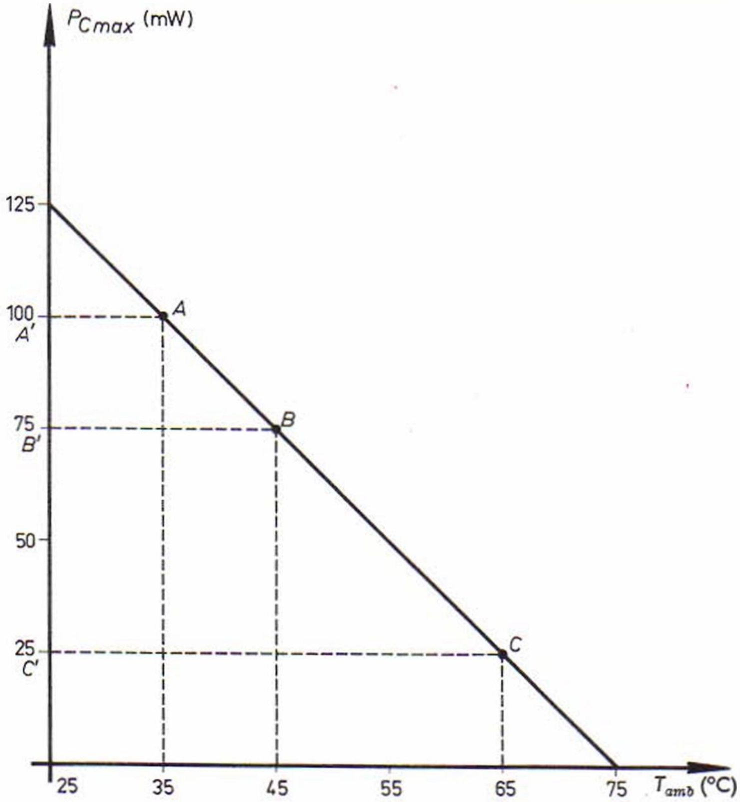


Fig. 154

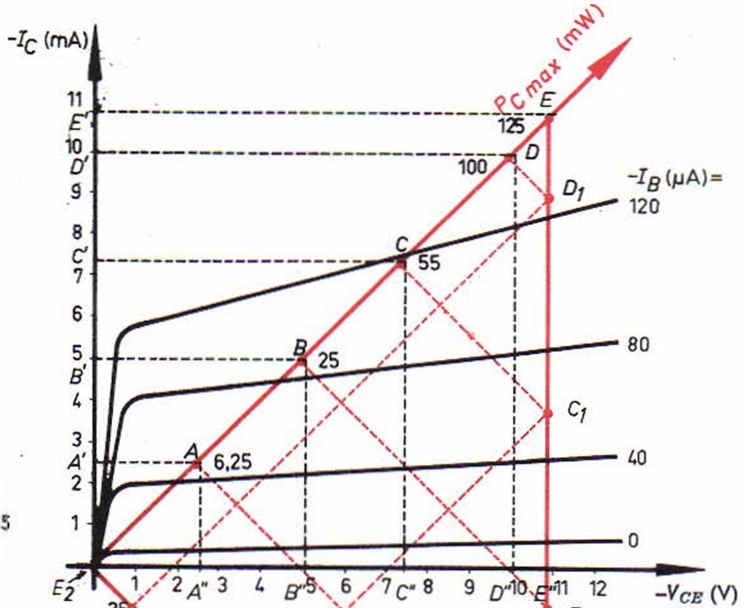


Fig. 155

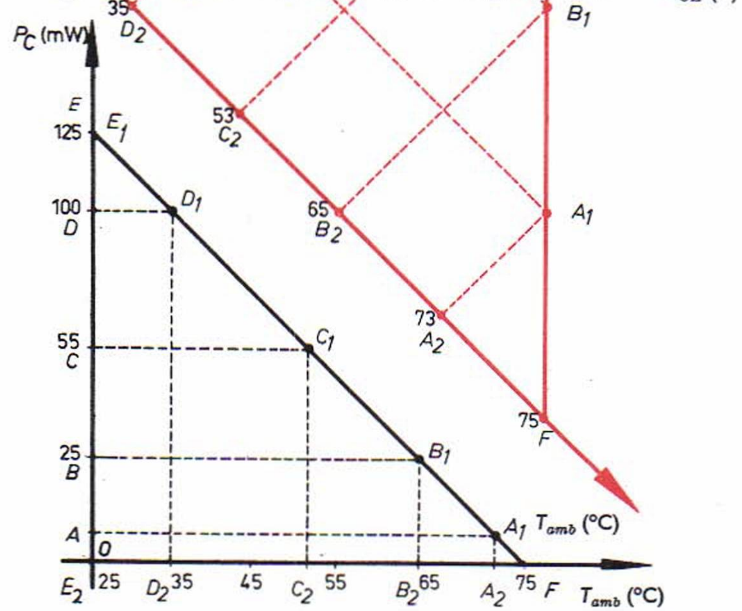


Fig. 156

The relationship between the characteristic and the working point of the transistor

We will start with the $-I_C = f(-V_{CE})$ characteristic in Fig. 155. The red ordinate $P_{C\max}$ is at an angle of 45° to the abscissa $-V_{CE}$. The scale for $-I_C$ is in milliamperes and that for $-V_{CE}$ in volts. The power scale on the $P_{C\max}$ axis is obtained by projecting A, B, C, D and E on the $-I_C$ axis (points A', B', C', D' and E') and on the $-V_{CE}$ axis (points A'', B'', C'', D'', and E''). The points A', B', C', D' and E' correspond to collector currents of 2.5, 5, 7.5, 10 and 11 mA, while points A'', B'', C'', D'' and E'' correspond to collector-emitter voltages of 2.5, 5, 7.5, 10 and 11 V.

Since the power dissipated in the collector is equal to the product of the collector-emitter voltage and the collector current,

$$P_C = V_{CE} \cdot I_C,$$

the collector dissipations at points A, B, C, D and E are 6.25, 25, 55, 100 and 125 mW respectively.

Points A, B, C, D and E correspond to points A_1, B_1, C_1, D_1 and E_1 on the $P_C = f(T_{amb})$ characteristic. The projections of the latter points on the T_{amb} axis determine the points A_2, B_2, C_2, D_2 and E_2. (Point F is found by marking off the distance $OE = 125$ mW on the T_{amb} axis, so that $OF = OE$. The $P_{C\max} = f(T_{amb})$ characteristic of Fig. 156 shows that a dissipation of 6.25 mW (A) is permissible at an ambient temperature of 73° C (A_2), a power of 25 mW (B) at an ambient temperature of 65° C (B_2), a power of 55 mW (C) at an ambient temperature of 53° C (C_2), a power of 100 mW (D) at an ambient temperature of 35° C (D_2) and a power of 125 mW (E) at an ambient temperature of 25° C (E_2).

Suppose that the transistor is to operate at an ambient temperature of 53° C, corresponding to point *E* (printed red) on the $P_{C\max} = f(T_{\text{amb}})$ characteristic of Fig. 157. The projection of point *E* on the $P_{C\max}$ axis corresponds to a maximum permissible collector dissipation of 55 mW. It is now possible to draw, through this point on the $-I_C = f(-V_{CE})$ characteristic, the hyperbola which is the locus of all points corresponding to a power of 55 mW.

In the red hatched region *A*, the collector dissipation is greater than 55 mW, and care must be taken to ensure that the load line remains outside this region under all circumstances.*

Region *B* corresponds to collector dissipations of less than 55 mW. The red load line of Fig. 157 can be used without any danger of the transistor being damaged.

If the transistor is to be used at a temperature of 65° C, the maximum permissible collector dissipation is only 25 mW. The hyperbola for this power (printed green in Fig. 158) is the limit of the green-hatched region outside which the load line must lie. (region *C*).

For $T_{\text{amb}} = 65°$ C the continuous green load line determines the load resistance at which the collector dissipation is a maximum, without the transistor being in danger of being damaged (region *D*). The broken green line is the load line which can be employed at an ambient temperature of 65° C under certain special conditions. Under these conditions the collector dissipation does in fact become greater than 25 mW if the working point shifts from *L* to *M*.

In fact, if the permissible load is a resistance, it depends on the ambient temperature. In order to establish an accurate relationship between the dissipation and the ambient temperature, we must take the following factors into account:

> The thermal conductivity from the "centre" of the transistor to its surroundings.
> The ambient temperature.
> Variations in the electrical power as a time function.

---

* This condition does not apply to a class *B* push-pull stage or to pulse drive.

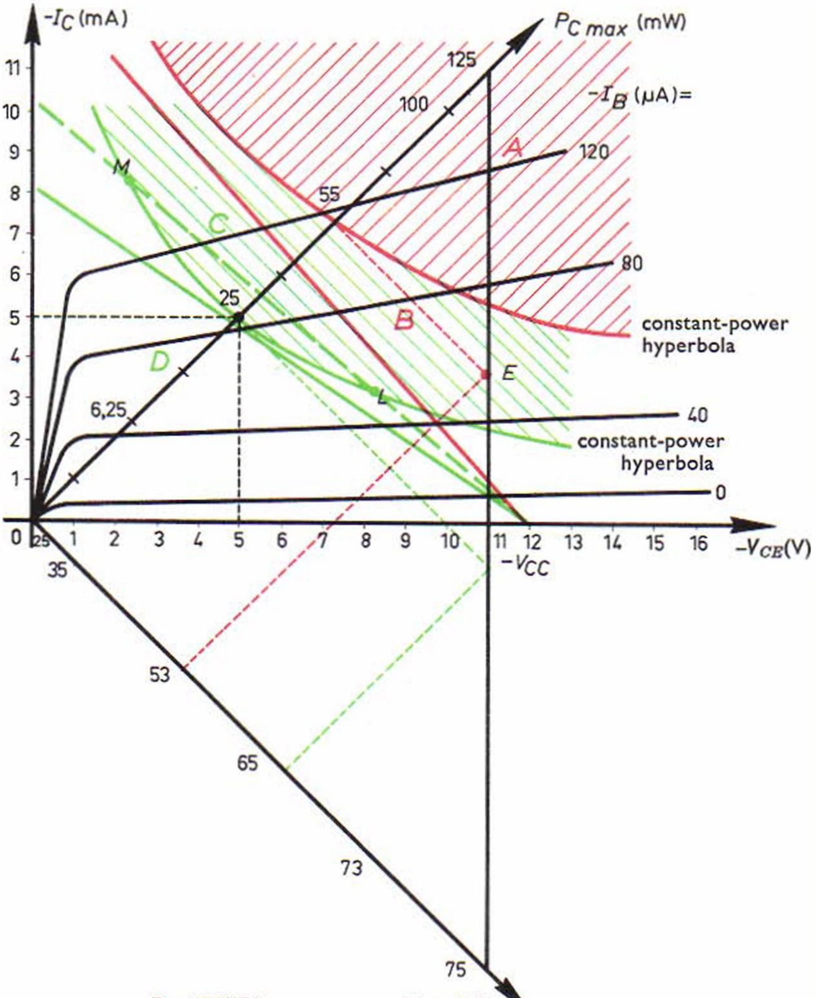

-I_C (mA)
$P_{C\,max}$ (mW)
-I_B (μA)=
120
80
40
0
125
100
55
25
6,25
A
B
C
D
E
L
M
constant-power hyperbola
constant-power hyperbola
-V_{CE}(V)
-V_{CC}
0
1
2
3
4
5
6
7
8
9
10
11
12
13
14
15
16
25
35
53
65
73
75
T_{amb} (°C)

Fig. 157/158

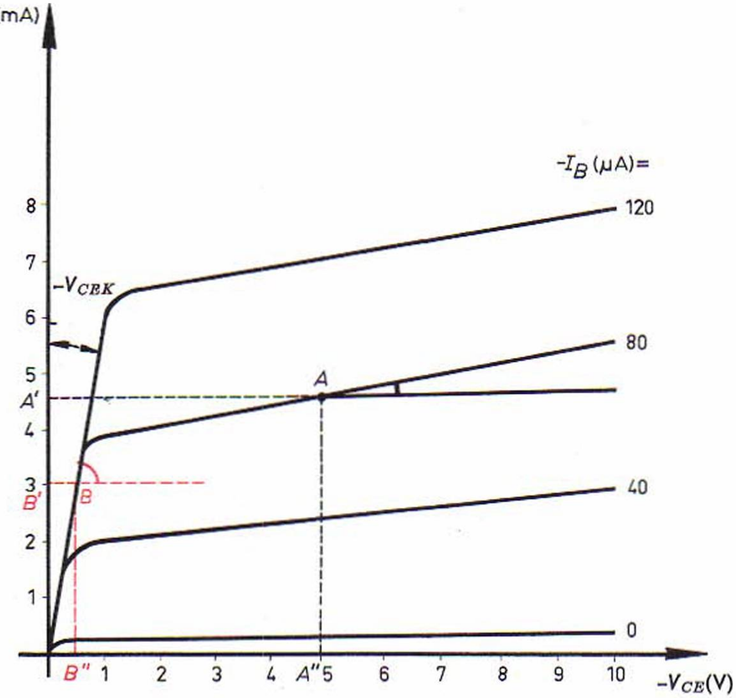


Fig. 159

f) Closer examination of the $-I_C = f(-V_{CE})$ characteristic

The $-I_C = f(-V_{CE})$ characteristic of Fig. 159 suggests that we might examine the following two points more closely:

> The collector-emitter voltage $-V_{CEK}$ at which the characteristic flattens out (the knee voltage)
> The leakage current $-I_{CEO}$.

Knee voltage

At very low voltages the characteristic shows a sharp bend, similar to that of the $I_a = f(V_a)$ characteristic of pentodes, but for transistors this bend occurs at much lower voltages than it does for pentodes. This is a very important advantage of transistors, because they can be driven so far that the collector-emitter voltage becomes very low.

Let us consider point A in this simplified and idealized family of characteristics; this point corresponds to point A'' on the abscissa and point A' on the ordinate. If we draw a horizontal line through point A, the angle between this line and the characteristic represents the output impedance of the transistor:

$$\tan A = 1/\mathrm{h}_o \text{ (compare page 137).}$$

Angle A is small, so that tan A is also small, and the output impedance h_o is therefore high.

Point B on the characteristic corresponds to B'' on the abscissa (a value lower than $-V_{CEK}$) and to a point B' on the ordinate. If we now draw a horizontal line through point B, this line makes a very large angle with the characteristic; tan B is now large, and the output impedance h_o is low. The knee voltage of a transistor is defined as the collector-emitter voltage at which, for a given collector current, the output impedance is subject to a sudden change. At collector-emitter voltages lower than $-V_{CEK}$ the output impedance of the transistor is very low, and only a little power is dissipated in the transistor itself.

Leakage current $-I_{CEO}$

The occurrence of the leakage current $-I_{CEO}$, which has already been explained on page 117, will now be examined in more detail on the basis of Fig. 160, in which the collector current is again plotted as a function of the collector-emitter voltage. This current varies considerably with the temperature, and this factor must be taken into account in the design of a circuit. The straight line in the figure is the load line, and the working point is at *A*. The variations of the input signal cause the collector current to swing to and fro about *A*. The amplitude of this displacement is limited, on one side by the knee voltage $-V_{CEK}$ and on the other side by the leakage current $-I_{CEO}$.

If the temperature increases, the leakage current will also increase (printed red in the figure). This increase in $-I_{CEO}$ causes point *A* to move in the direction of increasing current.

The correct choice of the working point is therefore of the greatest importance; the operating conditions of the transistor must always be chosen so that the chance of distortion occurring is as small as possible, particularly with large signals. With small signals the choice of the working point is less critical, and consequently this point is not in the first place determined by the leakage current $-I_{CEO}$.

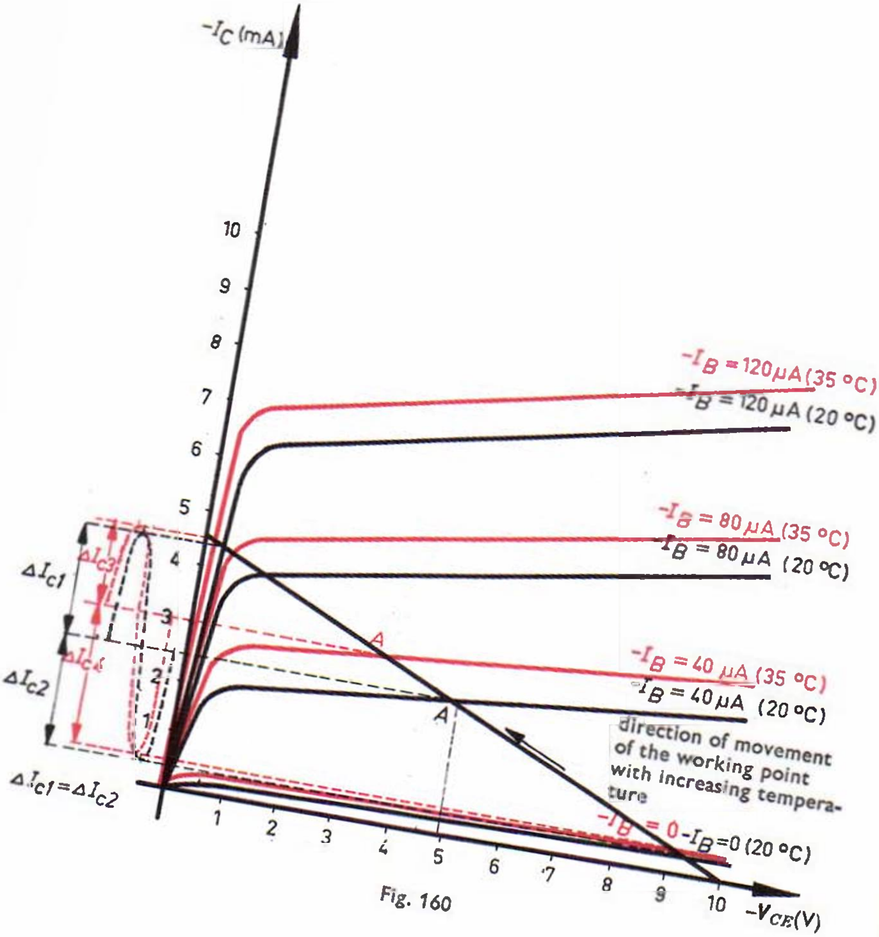

$-I_C$ (mA)
10
9
8
7
6
5
4
3
2
1
$-I_B = 120 \mu A$ (35 °C)
$-I_B = 120 \mu A$ (20 °C)
$-I_B = 80 \mu A$ (35 °C)
$-I_B = 80 \mu A$ (20 °C)
$-I_B = 40 \mu A$ (35 °C)
$-I_B = 40 \mu A$ (20 °C)
A
A
direction of movement of the working point with increasing temperature
$-I_B = 0$
$-I_B = 0$ (20 °C)
ΔI_{c1}
ΔI_{c2}
ΔI_{c3}
ΔI_{c4}
$\Delta I_{c1} = \Delta I_{c2}$
1
2
3
4
5
6
7
8
9
10
$-V_{CE}$(V)

Fig. 160

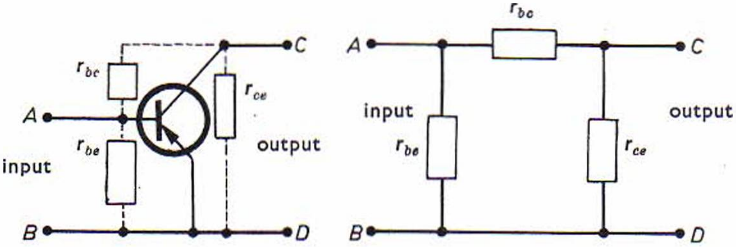

r_{bc}
A
r_{be}
input
B
r_{ce}
output
C
D
r_{bc}
A
C
input
r_{be}
output
r_{ce}
B
D

Fig. 161

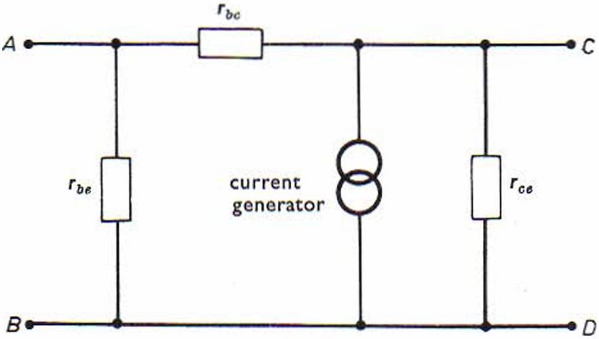

r_{bc}
A
C
r_{be}
current
generator
r_{ce}
B
D

Fig. 162

CHAPTER 20

Transistor parameters

The above discussion of transistor characteristics has shown that, in using these characteristics, account must be taken of the following important parameters:

The input impedance
The output impedance
The current gain
The internal feedback

The object of this chapter is to define these four fundamental parameters in more detail, and to study the variations which occur in them.

20.1. The equivalent circuit of the transistor

Fig. 161 represents the equivalent circuit of the transistor. Between the base and the emitter there is a resistance R_{be}, which is in fact the base-emitter resistance of the transistor. Between the base and the collector there is a resistance R_{bc}, representing the base-collector resistance or the feedback impedance of the transistor. Between the collector and the emitter there is a resistance R_{ce}, representing the collector-emitter resistance of the transistor.

In this way, we have defined the transistor as a passive element (for low frequencies), but we must also take its active operation into account. To do this, we connect a current generator across the output terminals, the internal resistance of this generator being infinite (see Fig. 162). The various parameters which we have just discussed are determined in principal by the mutual relationship between the electrical quantities.

The four parameters can be defined with the aid of the circuit of Fig. 163. In this figure v_1 represents the input voltage of the circuit, i_1 represents the current flowing in the input circuit, v_2 the voltage across the output terminals, and i_2 the current flowing in the output circuit. The quantities associated with the input circuit are indicated by the subscript 1 and those associated with the output circuit by the subscript 2.

It is now possible to determine the mutual relationships between these four quantities (i_1, i_2, v_1 and v_2). The relationship between two of these quantities is represented by the symbol h with a letter subscript. (Previously, two figure-subscripts were used, corresponding to the subscripts appearing in the numerator and the denominator).

The relationship between the output current and the input current is thus represented by the symbol:

$$\mathrm{h}_f (= \mathrm{h}_{21}) = i_2/i_1. \qquad (f \text{ for "forward"})$$

In defining these parameters it is also important to know how the transistor is connected. For this reason therefore, it is also necessary to indicate whether the transistor is in common base, common emitter, or common collector. To this end, the symbol h_f is followed by a second subscript (*b*, *e* or *c*), the small letter indicating which electrode is common to both input and output circuits. In the common-emitter configuration, therefore, the ratio of the output current to the input current is represented by the symbol h_{fe}.

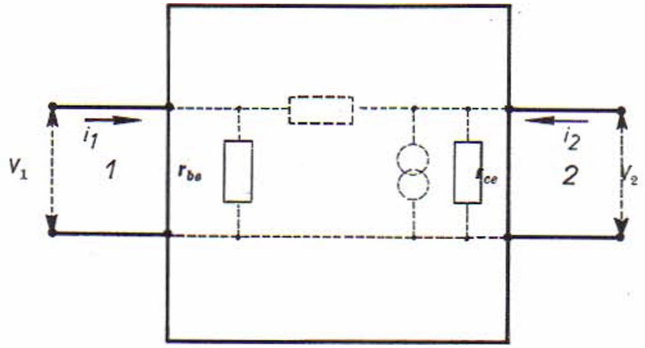

i_1
i_2
V_1
1
r_{be}
r_{ce}
2
V_2

Fig. 163

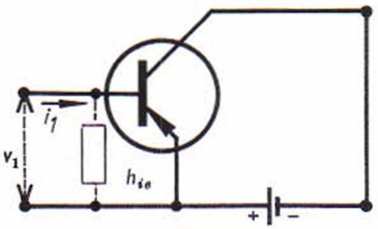


Fig. 164 a

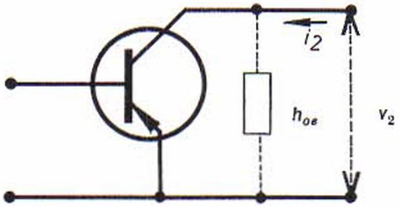


Fig. 164 b

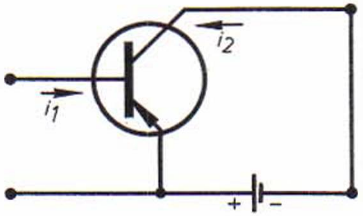


Fig. 164 c

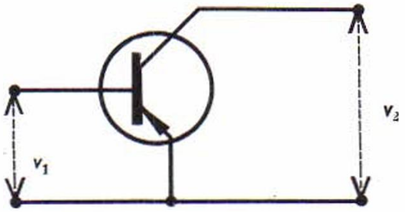


Fig. 164 d

The ratio of the input voltage v_1 to the input current i_1 with short-circuited output (Fig. 164a) is known as the input impedance of the transistor, and is represented by:

$$h_{ie}\ (= h_{11e}) = v_1/i_1 \qquad (i \text{ for "input"})$$

The ratio of the output voltage v_2 to the current through the output circuit i_2 with open-circuited input (Fig. 164b) is termed the output impedance of the transistor. However, it is more usual to employ the reciprocal of this quantity, i.e. the output admittance of the transistor:

$$h_{oe}(= h_{22e}) = i_2/v_2. \qquad (o \text{ for "output"})$$

The ratio of the output current i_2 to the input current i_1 with short-circuited output (Fig. 164c) is termed the current gain of the transistor:

$$h_{fe} = (h_{21e}) = i_2/i_1. \qquad (f \text{ for "forward"})$$

The ratio of the input voltage v_1 to the output voltage v_2, with open-circuited input (Fig. 164d) is termed the internal feedback of the transistor.

$$h_{re}\ (= h_{12e}) = v_1/v_2. \qquad (r \text{ for "reverse"})$$

It should be remembered that these parameters may only be used for small signals and under special conditions (either open-circuited input or short-circuited output).

We will now examine the methods which enable us to measure and investigate the variations experienced by the parameters which determine the electrical behaviour of the transistor (input impedance, output impedance, current gain and internal feedback).

20.2. The input impedance of the transistor

The input impedance is the ratio of the input voltage, (base-emittervoltage) to the current flowing in the base circuit.

a) Measurement of the input impedance

To measure the input impedance we can use the circuit shown in Fig. 165.

Switch 1 open

The a.f. generator which is connected across terminals *A* and *B* supplies a voltage of 10 V for example, at a frequency of 1 kc/s. The current flowing in the input circuit (Fig. 166) depends on the resistance (in this case 1 MΩ) in series with the generator and on the impedance of the transistor, which can be neglected here. The current flowing in the input circuit is thus:

$$i_b = 10/10^6 = 10^{-5}\text{ A} = 10\ \mu\text{A}.$$

Suppose that the current gain $h_{fe} = 50$. The collector current will then be equal to:

$$i_c = h_{fe}.i_b = 50 \times 10 = 500\ \mu\text{A}.$$

The voltage across the load resistance of 1 kΩ in the collector circuit is equal to:

$$V_{RL} = R_L.i_c = 10^3 \times 5 \times 10^{-4} = 0.5\text{ V}.$$

This voltage can be measured by means of a valve voltmeter across terminals *C* and *D*.

Switch 1 closed

Let us now examine the situation when switch 1 is closed (Fig. 167). The equivalent circuit is drawn in Fig. 168. Potentiometer R_1 is adjusted so that the valve voltmeter now indicates half its previous reading (0.25 V). Consequently, the collector current is now equal to:

$$i_c = v_{RL}/R_L = 0.25/10^3 = 25 \times 10^{-5}\text{ A} = 250\ \mu\text{A}.$$

which corresponds to a base current of:

$$i_b = i_c/h_{fe} = 250/50 = 5\mu\text{ A}.$$

The current flowing through the 1 MΩ resistance is equal to the sum of the base current and the current through the potentiometer. The current *i* is still 10 μA, since h_i and R_1 can be neglected in relation to 1 MΩ. From this we see that $i_{R_1} = i - i_{h_i} = 10 - 5 = 5\ \mu\text{A}$.

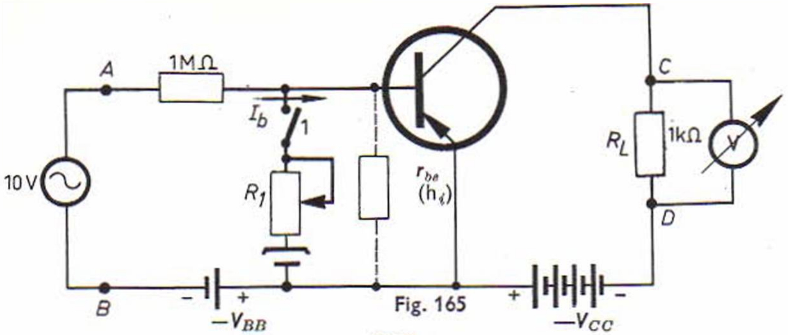

A
1MΩ
I_b
1
R_1
10V
r_{be}
(h_i)
C
R_L
1kΩ
V
D
B
– +
$-V_{BB}$
+ –
$-V_{CC}$

Fig. 165

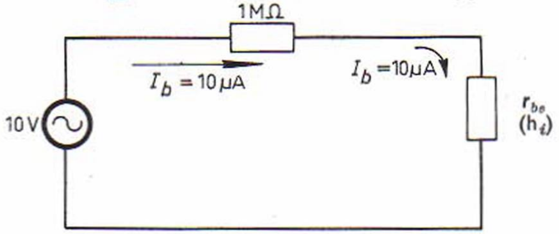

1MΩ
I_b = 10 µA
I_b =10µA
10V
r_{be}
(h_i)

Fig. 166

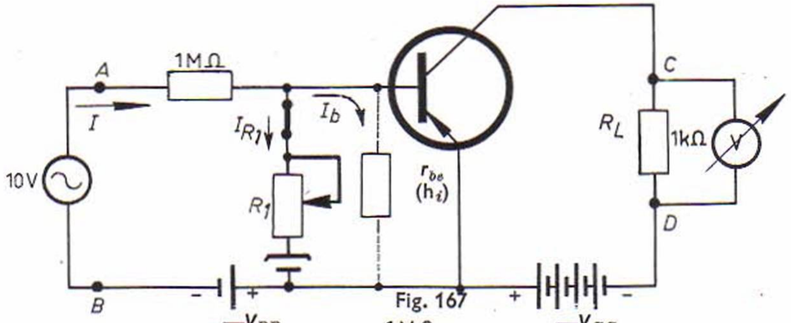

A
1MΩ
I
I_{R1}
I_b
R_1
10V
r_{be}
(h_i)
C
R_L
1kΩ
V
D
B
– +
$-V_{BB}$
+ –
$-V_{CC}$

Fig. 167

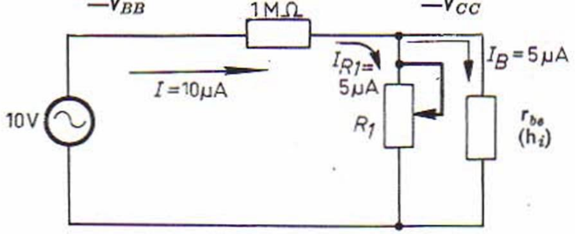

1MΩ
I =10µA
I_{R1}= 5µA
I_B = 5µA
R_1
10V
r_{be}
(h_i)

Fig. 168

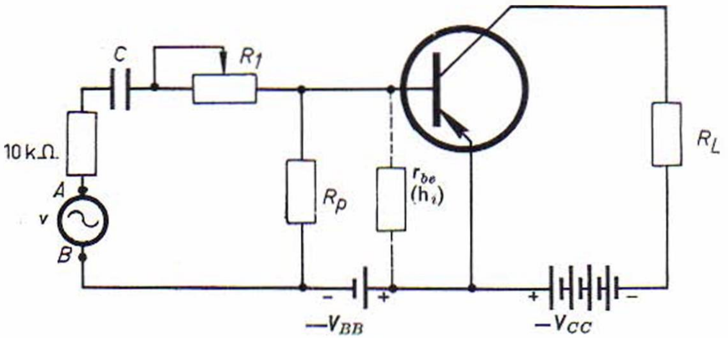

C
R1
10 kΩ
A
v
B
Rp
r_be
(h_i)
R_L
−
+
−V_BB
+
−
−V_CC

Fig. 169

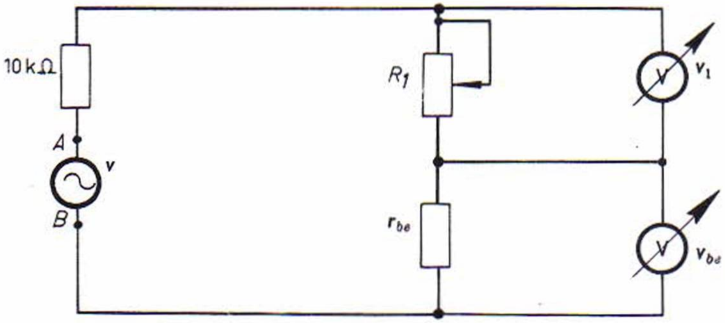

10 kΩ
A
v
B
R1
r_be
V
v1
V
v_be

Fig. 170

As the currents i_{h_i} and i_{R1} are equal, the input impedances h_i and R_1 are also equal. The input impedance of the transistor can thus be determined by measuring the value of R_1 by means of an ohmmeter. However, this measurement is only valid if the $-I_C = f(-I_B)$ characteristic is completely linear. (only then is the current gain constant).

As the method we have just described can only be used in certain cases, it is desirable to have a more general method of determining the input impedance of a transistor for large powers. This is because a transistor of this type has a sharply curved $-I_C = f(-I_B)$ characteristic. The required method is based on the circuit of Fig. 169.

The a.f. generator is connected across terminals *A* and *B*, and drives the transistor via a resistance of 10 kΩ and the variable resistance R_1. The decoupling capacitor *C* prevents the direct-current operating conditions of the transistor from being disturbed by the a.f. generator. Valve voltmeters for reading the a.f. alternating voltage are connected across the potentiometer and between the base and emitter of the transistor.

Fig. 170 represents the equivalent circuit. The potentiometer is adjusted until both voltmeters indicate the same voltage. The voltages across the resistance R_1 and the input impedance $r_{be} = h_i$ are then equal, ($v_1 = v_{be}$), since these are in series and the same current flows through both. The input impedance is thus equal to the value of R_1, which can be determined by means of an ohmmeter. For this method it is necessary to have two valve voltmeters, hence it will only be employed if the characteristic of the transistor is not linear.

b) Variation of the input impedance as a function of the collector current

It is possible to measure the input impedance of the transistor for various values of the collector current $-I_C$. This can be done by one of the methods already discussed, when it will be found that the curve $h_i = f(-I_C)$ is as shown in Fig. 171. From this we see that the input impedance decreases sharply as the collector current increases. The variations of the input impedance as a function of the base-emitter voltage, and thus of the base current, have already been explained on page 154. As a reminder, Fig. 145 is repeated here as Fig. 172.

c) Variation of the input impedance as a function of the collector-emitter voltage

In order to determine the variation of the input impedance as a function of the collector-emitter voltage we employ the circuit of Fig. 173. By means of potentiometer R_1 the collector-emitter voltage can be adjusted to various values, which can be read from the d.c. voltmeter. The input impedance is found to increase slightly with the collector-emitter voltage, as shown by the curve of Fig. 174.

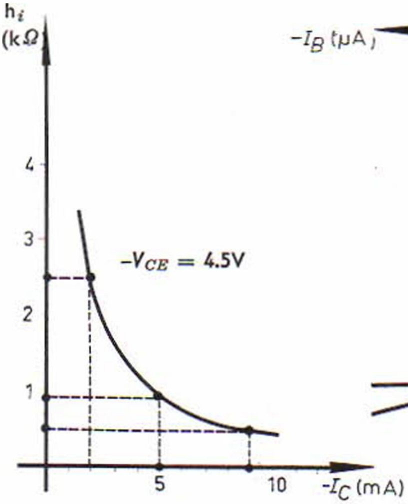


Fig. 171

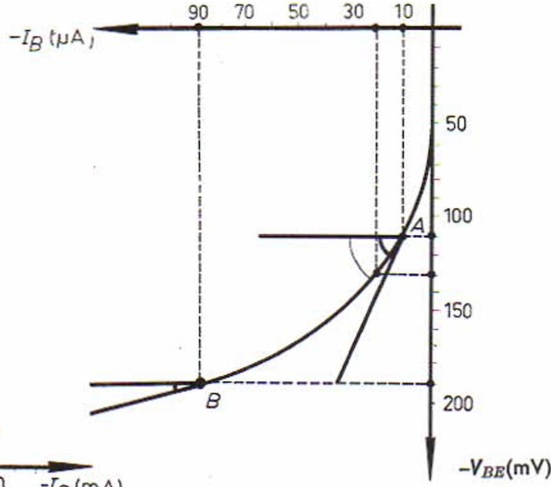


Fig. 172

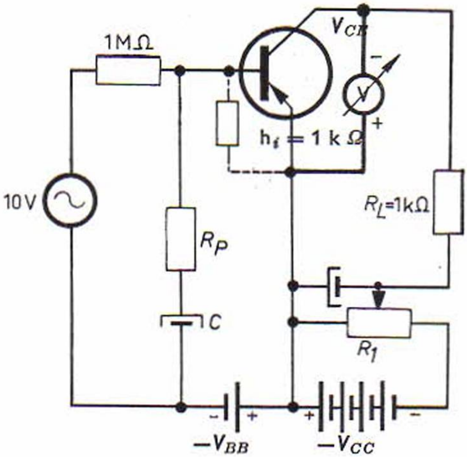

1MΩ
V_CE
V
-
+
h_i = 1 kΩ
10V
R_P
C
R_L=1kΩ
R_1
-
+
-V_BB
+
-
-V_CC

Fig. 173

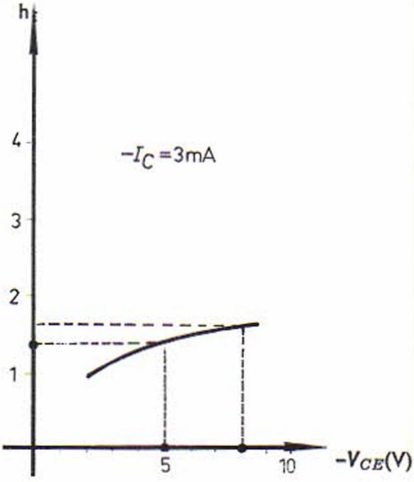


Fig. 174

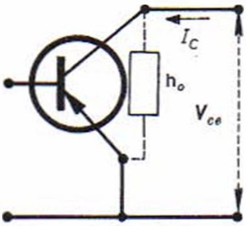
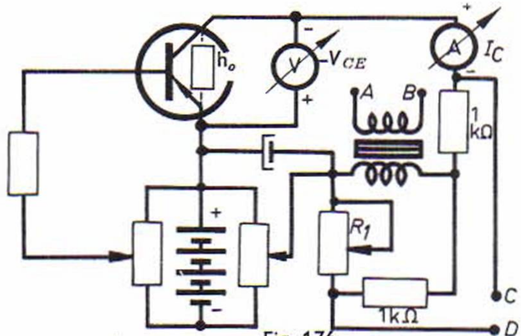
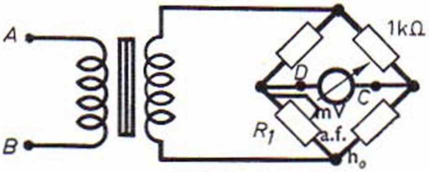


Fig. 175



h$_o$
V
$-V_{CE}$
A
I_C
+
-
A
B
1 kΩ
R_1
1kΩ
C
D

Fig. 176



A
B
1kΩ
D
C
mV
a.f.
R_1
h$_o$

Fig. 177

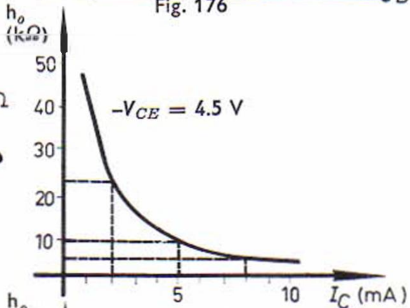


Fig. 178

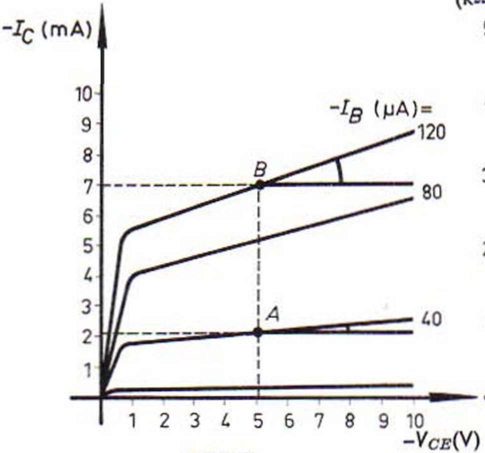


Fig. 179

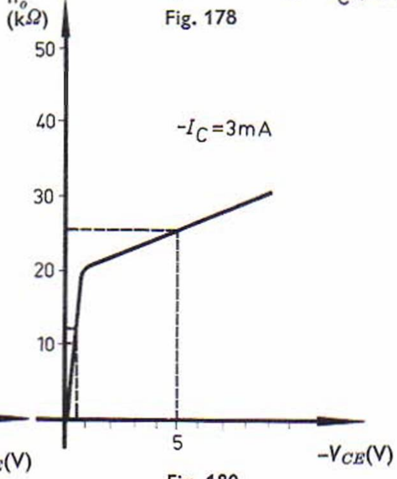


Fig. 180

20.3. The output impedance of the transistor

The output impedance of the transistor is the ratio of the collector-emitter voltage v_{ce} to the collector current i_c (see Fig. 175).

a) Measurement of the output impedance

The output impedance can be measured by means of the circuit shown in Fig. 176. The a.f. generator is connected across the terminals A and B and the a.f. voltmeter is connected between terminals C and D. Fig. 177 is the corresponding equivalent circuit. The bridge circuit formed in this way is in equilibrium if the resistances h_o and R_1 are equal, in which case the voltage between terminals C and D is a minimum. This state of equilibrium can be adjusted by means of potentiometer R_1. The output impedance h_o of the transistor is equal to the value of R_1, which can be measured by means of an ohmmeter.

b) Variation of the output impedance as a function of the collector current

The method described here can be used to measure the output impedance of the transistor for various values of the collector current $-I_C$. This impedance is found to decrease rapidly as the collector current increases, as shown in Fig. 178. This can be explained on the basis of the $-I_C = f(-V_{CE})$ characteristic. (see Fig. 179).

The angle between this characteristic and the horizontal defines the output impedance, according to the equation:

$$\tan A = 1/h_o.$$

This angle increases with the collector current, so that its tangent also increases, and the output impedance thus decreases. Fig. 180 shows how the output impedance of a transistor decreases gradually with the collector-emitter voltage, until the latter drops below the knee voltage $-V_{CEK}$; past this point the output impedance drops very rapidly to a very low value.

20.4. The current gain of the transistor

The current gain h_{fe} of the transistor is defined as the ratio of the collector current i_c to the base current i_b in the grounded-emitter configuration (Fig. 181).

a) Measurement of the current gain

We will assume that the transistor has a practically linear $-I_C = f(-I_B)$ characteristic, like type OC 71, for example. To measure the current gain, we employ the circuit of Fig. 182. The a.f. generator is connected to the base-emitter circuit via a resistance of 1 MΩ. The decoupling capacitor prevents the direct-current operating conditions from being disturbed by the a.f. generator. The equivalent circuit is illustrated in Fig. 183. If the a.f. generator has an output voltage of 10 V, the resulting current flowing in the base-emitter circuit is:

$$i_b = 10/(10^6 + h_i) \backsimeq 10^{-5}\ \text{A} = 10\ \mu\text{A}.$$

since h_i can be neglected in relation to the resistance of 1 MΩ.

The a.f. voltmeter which is connected across the load resistance of 1 kΩ indicates the voltage across this resistance. Suppose that this voltage $V_o = 0.5$ V. This means that the collector current is:

$$i_c = V_o/R_L = 0.5/10^3 = 5 \times 10^{-4}\ \text{A} = 500\ \mu\text{A}.$$

The current gain of the transistor can now be calculated from the known values of the collector current i_c and the base current i_b:

$$h_{fe} = i_c/i_b = 500/10 = 50.$$

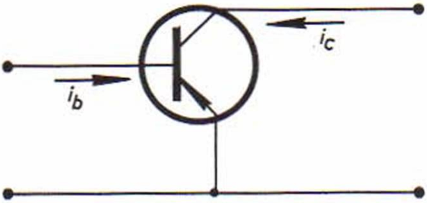


Fig. 181

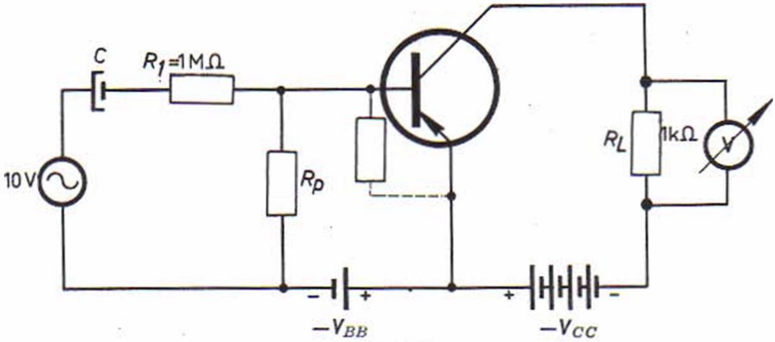

C
R_1=1MΩ
10V
R_p
R_L
1kΩ
V
- +
-V_BB
+ -
-V_CC

Fig. 182

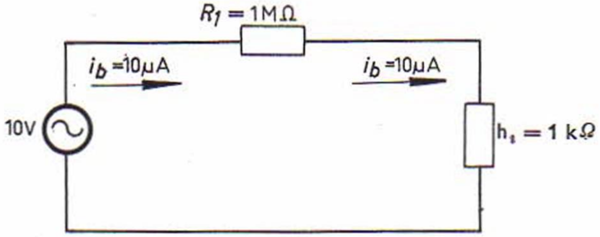

R_1 = 1MΩ
i_b =10µA
i_b =10µA
10V
h_i = 1 kΩ

Fig. 183

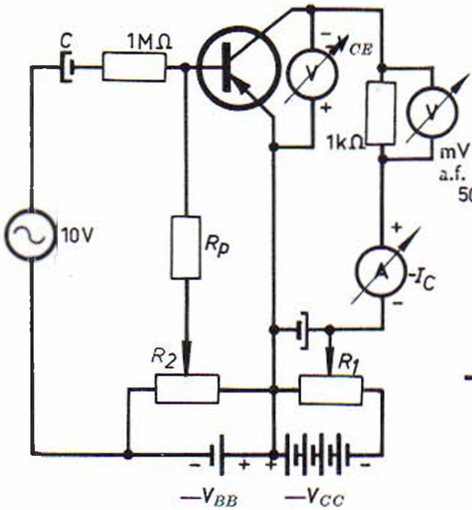

C
1MΩ
−V_{CE}
+
V
1kΩ
V
mV
a.f.
10V
R_p
+
A
$-I_C$
−
R_2
R_1
− + + −
$-V_{BB}$
$-V_{CC}$

Fig. 184

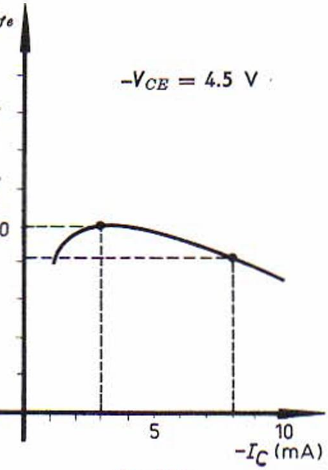


Fig. 185

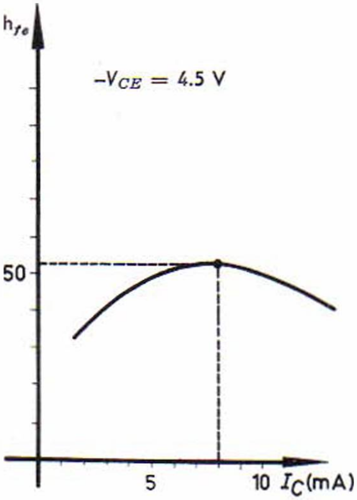


Fig. 186

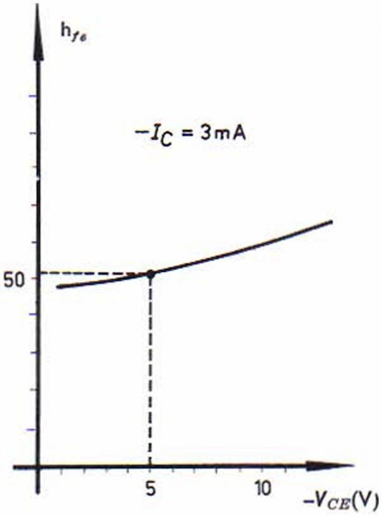


Fig. 187

b) Variation of the current gain as a function of the collector current

To determine the variation of the current gain as a function of the collector current, we can use the circuit of Fig. 184. The base-emitter voltage $-V_{BE}$, and thus the base current $-I_B$ can be adjusted by means of R_2. Variations in the base current will cause variations in the collector current $-I_C$. The current gain can now be measured as already described, for various values of the collector current, which can be measured by means of the milliammeter connected in series with the load resistance R_L. With an a.f. generator output voltage of 10 V, the current in the base-emitter circuit of the transistor will remain equal to 10 μA because, although the transistor input impedance will vary with the collector current, this impedance will always be negligible in relation to the 1 MΩ resistance.

For a small-signal transistor, such as the OC 71, this curve will be practically flat, as can be seen from Fig. 185. For a transistor for higher powers, whose $-I_C = f(-I_B)$ characteristic is not linear, however, the current gain will decrease with increasing collector current (see Fig. 186).

c) Variation of the current gain as a function of the collector-emitter voltage

In order to measure the variation of the current gain as a function of the collector-emitter voltage, we connect a d.c. voltmeter between the collector and the emitter of the transistor. The collector-emitter voltage can be adjusted by means of R_1. The current gain can now be determined as described above, for various values of the collector-emitter voltage, giving a curve as shown in Fig. 187. This shows that the current gain is practically independent of the collector-emitter voltage.

20.5. The internal feedback of the transistor

The internal feedback of the transistor is defined as the ratio of the base-emitter voltage v_{be} to the collector-emitter voltage v_{ce}.

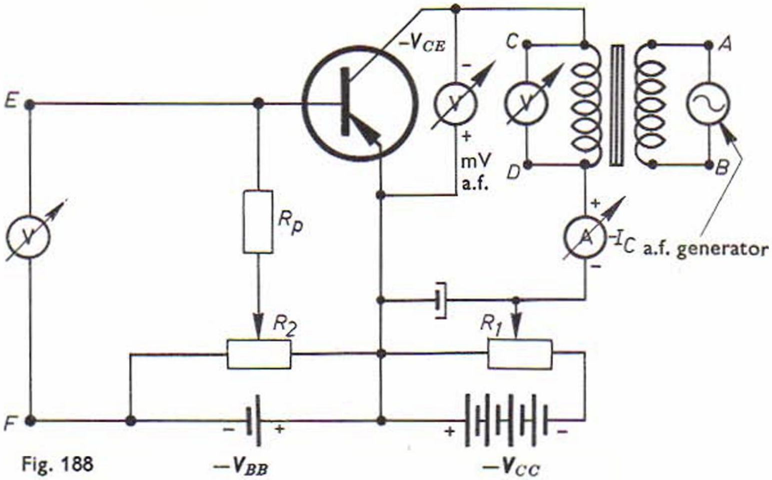
a) Measurement of the internal feedback

The internal feedback of a transistor can be measured by means of the circuit shown in Fig. 188. The a.f. generator is connected to terminals *A* and *B*, and the a.f. valve voltmeter to terminals *C* and *D*. The a.f. generator is adjusted so that this meter indicates a voltage of 0.5 V. With the same voltmeter, we now measure the voltage between the terminals *E* and *F*. This is the base-emitter voltage v_{be} of the transistor. The internal feedback is $h_{re} = v_{be}/v_{ce}$, and is also given by the ratio of the input impedance of the transistor to its output impedance. The equivalent circuit, including the internal feedback, is shown in Fig. 189.

b) Variation of the internal feedback as a function of the collector current

Since the collector current $-I_C$ is determined principally by the base current, the former depends on the position of potentiometer R_2. We can now repeat the above measurements for different values of the collector current, this current being measured by means of the d.c. ammeter connected in series with the load resistance.

The curve obtained in this way will be similar to the one shown in Fig. 190. From this we see that the feedback decreases sharply with increasing current, for low values of the collector current, but remains practically constant at higher values of the current. The explanation of this effect is obvious, since the internal feedback of the transistor is defined as the ratio of the input impedance to the output impedance, and the former decreases as the collector current increases.



E
F
$-V_{CE}$
C
D
A
B
mV
a.f.
R_p
R_2
R_1
$-I_C$
a.f. generator
$-V_{BB}$
$-V_{CC}$

Fig. 188

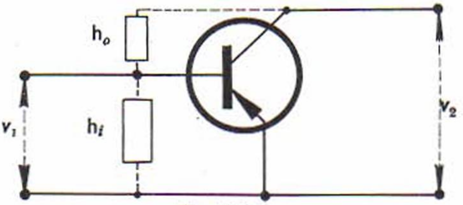


Fig. 189

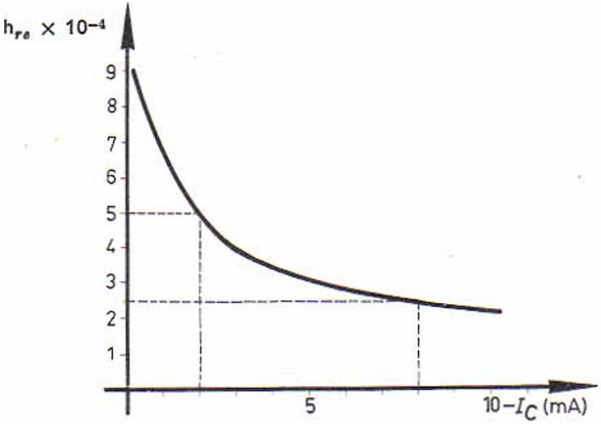


Fig. 190

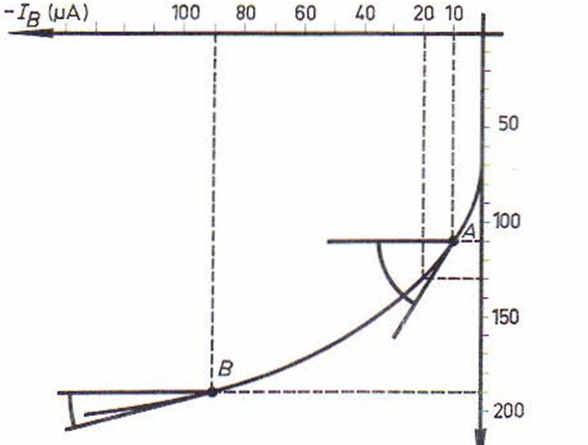


Fig. 191

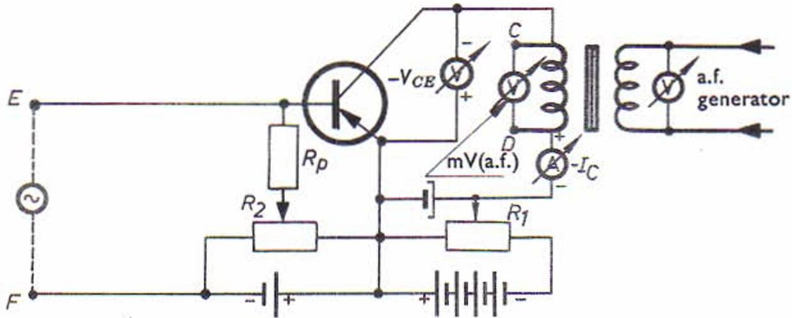

E
F
R_p
R_2
R_1
-V_{CE}
V
C
D
mV(a.f.)
A
-I_C
a.f.
generator
$-V_{BB}$
$-V_{CC}$

Fig. 192

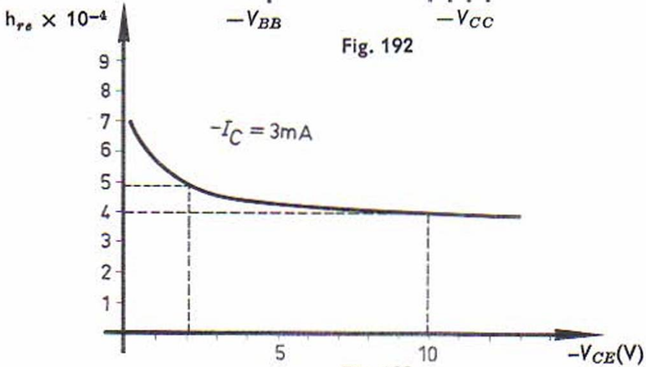


Fig. 193

The $-I_B = f(-V_{BE})$ characteristic of Fig. 191 shows that the input impedance is high at low values of the base current, and low for high values of this current; at very low values of the base current the input impedance increases steeply. This effect has already been mentioned on page 154 in connection with the variation of the input impedance as a function of the collector current.

The internal feedback increases with the value of the input impedance, and is a maximum at a low base current (low collector current); at higher values of the base current (and collector current) the feedback approaches a constant value.

c) Variation of the internal feedback as a function of the collector-emitter voltage

To measure the variation of the internal feedback as a function of the collector-emitter voltage, we use the circuit of Fig. 192. By means of potentiometer R_1 we can adjust the collector-emitter voltage, which is then measured by means of the d.c. voltmeter between collector and emitter. We now repeat the measurements which have already been described, for different values of this voltage. The resulting curve is as shown in Fig. 193, from which we see that the internal feedback of a transistor decreases slightly as the collector-emitter voltage increases.

Forward transfer admittance or slope

The four parameters which have been discussed in the previous chapter are much used for a.f. applications. For r.f. circuits, y parameters, of which the forward transfer admittance or slope of the transistor is the most important, are normally used. This is because it is possible to drive a transistor in a number of different ways, and with one of these (voltage drive) there are great advantages to be gained by introducting the conception of "slope", represented by the symbol y_{fe}.

21.1. Thermionic valves

The slope of a thermionic valve is given by its $I_a = f(V_g)$ characteristic, that is, by the curve which represents the anode current as a function of the grid current. (Fig. 194). This quantity is given by the expression:

$$S = \Delta I_a/\Delta V_g.$$

If we term the grid-cathode circuit of a valve the input circuit, and the anode-cathode circuit the output circuit (Fig. 195), we can describe the slope of a valve as the ratio of the variation of current in the output circuit to the voltage variation in the input circuit which causes this variation of current.

21.2. Transistors

We can start from the same basis idea with transistors. To this end, we term the base-emitter circuit the input circuit, and the collector-emitter circuit the output circuit (see Fig. 196). By the slope of a transistor, we understand the ratio of variations of the collector current (the current in the output circuit) to the corresponding variations of the base-emitter voltage (the voltage across the terminals of the input circuit):

$$y_{fe} = I_C/V_{BE}.$$

This slope can be deduced from the $-I_C = f(-V_{BE})$ characteristic of the transistor. This characteristic, which is often not supplied by the manufacturer, can easily be deduced from the $-I_C = f(-I_B)$ and $-I_B = f(-V_{BE})$ characteristics.

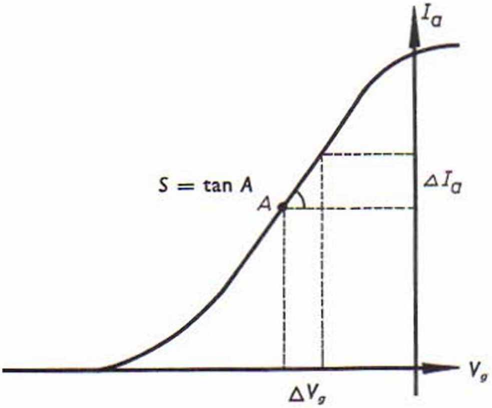


Fig. 194

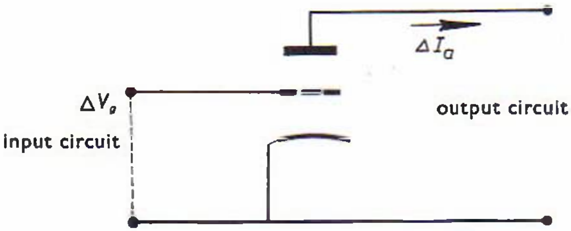

ΔI_a
ΔV_g
input circuit
output circuit

Fig. 195

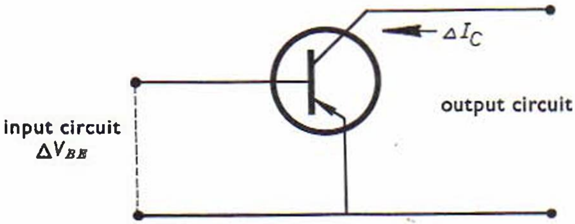

ΔI_C
output circuit
input circuit
ΔV_{BE}

Fig. 196

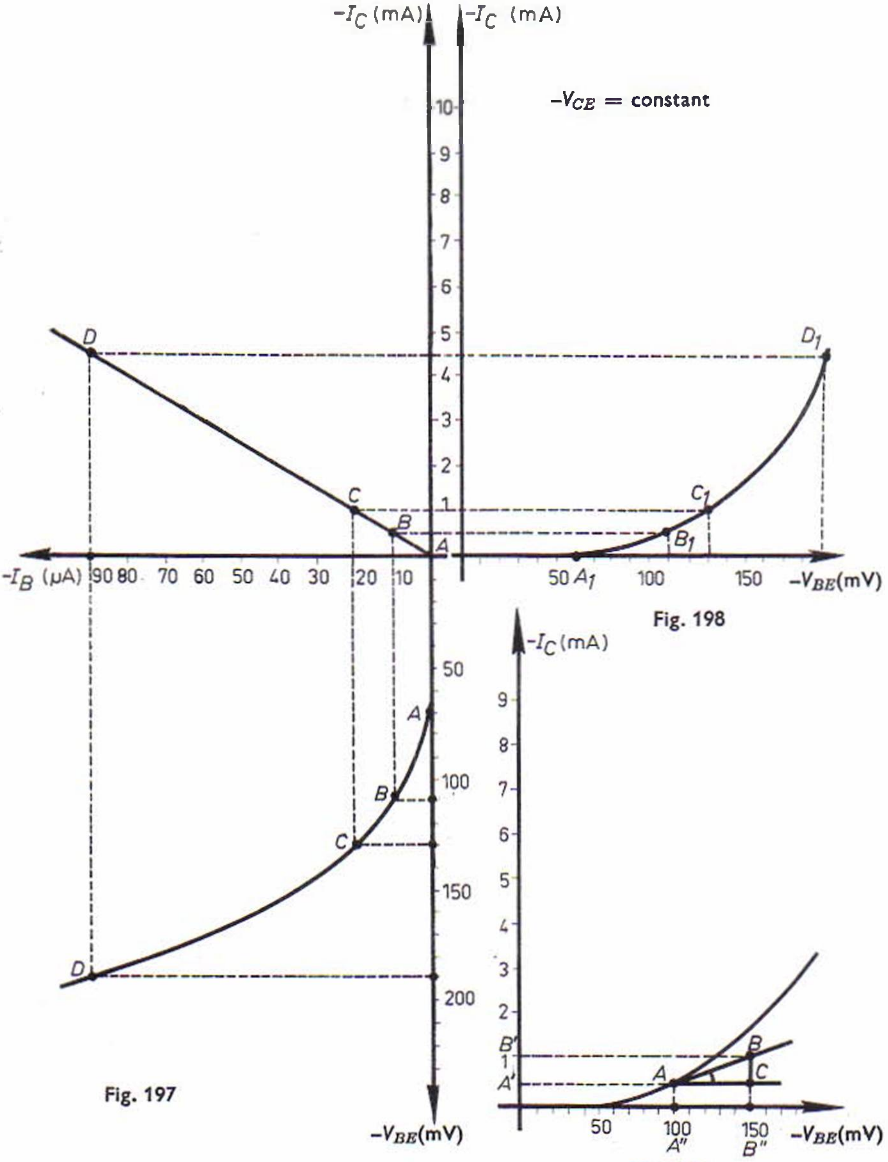


Fig. 198

Fig. 197

Fig. 199

21.3. Derivation of the $-I_C = f(-V_{BE})$ characteristic

In order to deduce the $-I_C = f(-V_{BE})$ characteristic of a transistor from the $-I_C = f(-I_B)$ and $-I_B = f(-V_{BE})$ characteristics given in Fig. 197, we set to work as follows. As shown in Fig. 198, we mark out $-V_{BE}$ along the abscissa, and $-I_C$ along the ordinate.

From Fig. 197 it follows that point A, with $-V_{BE} = 70$ mV, corresponds to $-I_B = 0$ and $-I_C = 0$. In Fig. 198 we now mark the point which corresponds to $-V_{BE} = 70$ mV and $-I_C = 0$ (point A_1). Similarly we find that the points $-I_C = 0.5$ mA, 1 mA and 4.5 mA (points B_1, C_1, and D_1) correspond to $-V_{BE} = 110$ mV, 130 mV and 190 mV.

Using the set of co-ordinates given in Fig. 197, therefore, it is sufficient to determine the collector current for a given value of V_{CE} as a function of the base-emitter voltage, and to plot the values thus obtained as in Fig. 198.

21.4. Determination of the slope from the $-I_C = f(-V_{BE})$ characteristic

In order to determine the slope of a transistor from the $-I_C = f(-V_{BE})$ characteristic derived as described above, we select a point A, at which $-V_{BE}$ is 100 mV (point A'') as shown in Fig. 199. Through this point we draw a horizontal line which gives us point A' on the $-I_C$ axis. We now draw the tangent to the curve at point A; this tangent forms an angle A with the horizontal. If we now take a point B'', such that $-V_{BE} = 150$ mV, this corresponds to point C on the horizontal line through point A, and to point B on the tangent to the curve through point A. The projection of point B onto the $-I_C$ axis gives point B'. The slope of the transistor is now:

$$y_{fe} = \tan A = BC/AC = A'B'/A''B'' = \Delta I_C/\Delta V_{BE}.$$

The tangent of the angle between the characteristic and the horizontal determines the slope of the transistor at a given point.

21.5. Variation of the slope as a function of the collector current

Let us consider point A on the characteristic shown in Fig. 200, at which $-V_{BE} = 100$ mV (Point A''). The horizontal line through this point gives point A' on the $-I_C$ axis. We also draw the tangent to the curve at point A. Point B'' on the abscissa, at which $-V_{BE} = 150$ mV, corresponds to point C on the horizontal line through point A and to point B on the tangent to the curve through this point. The projection of B on the $-I_C$ axis gives point B'.

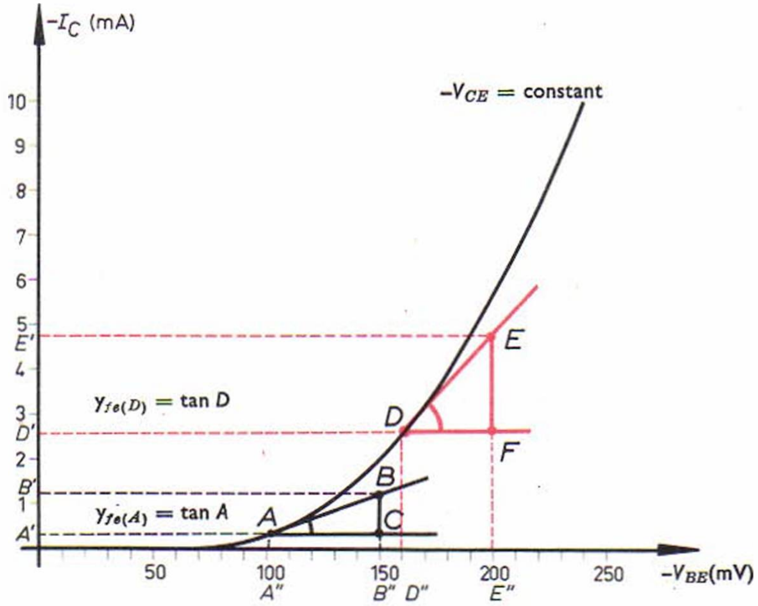
The slope of the curve at point A is now given by:

$$y_{fe(A)} = \tan A = BC/AC = A'B'/A''B'' = \Delta I_C/\Delta V_{BE} = 1/0.05 = 20 \text{ mA/V}.$$

Suppose we had taken point D'', corresponding to $-V_{BE} = 160$ mV, instead of point A''. This determines the working point D on the $-I_C = f(-V_{BE})$ characteristic. The horizontal line through point D (printed red) determines point D' on the $-I_C$ axis. If we now erect the perpendicular at point E'', corresponding to $-V_{BE} = 200$ mV, this gives us point F on the horizontal line through point D, and point E on the tangent to the curve at point D. The projection of point E on the $-I_C$ axis gives point E'. Consequently, the slope at point D is:

$$y_{fe(D)} = \tan D = FE/DF = D'E'/D''E'' = \Delta I_C/\Delta V_{BE} = 2.4/0.04 = 60 \text{mA/V}.$$

The above calculations show that the slope of the transistor increases with the collector current.



$-I_C$ (mA)
10
9
8
7
6
5
E'
4
3
D'
2
B'
1
A'
$-V_{CE}$ = constant
$y_{fe(D)}$ = tan D
$y_{fe(A)}$ = tan A
A
B
C
D
E
F
50
100
A''
150
B'' D''
200
E''
250
$-V_{BE}$(mV)

Fig. 200

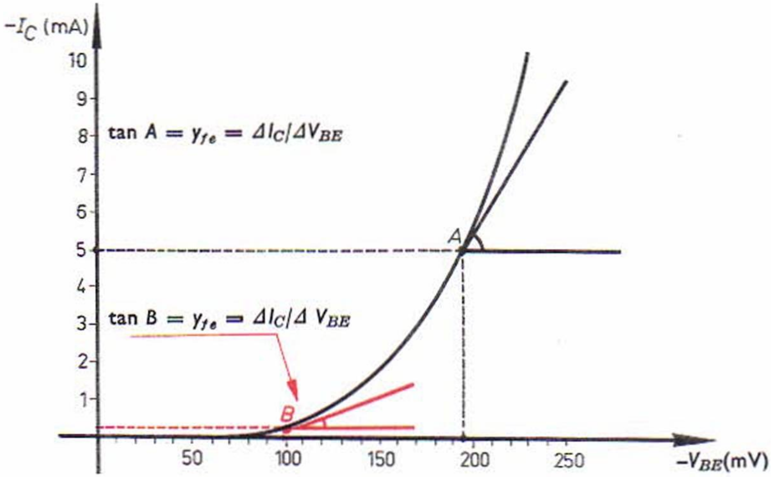
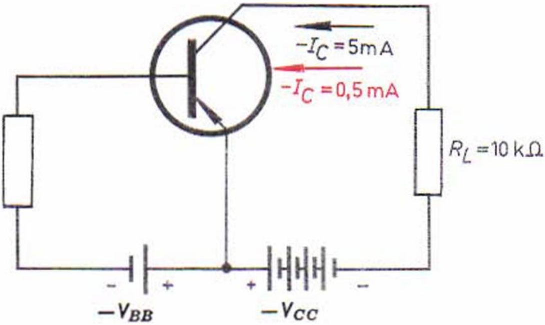


Fig. 201



$-I_C = 5\,mA$
$-I_C = 0{,}5\,mA$
$R_L = 10\,k\Omega$
-
+
+
-
$-V_{BB}$
$-V_{CC}$

Fig. 202

The gain of the transistor depends on its slope, provided that the signal conveyed to the input can be regarded as a voltage, and in that case a large slope is important in obtaining a large gain. To achieve this, however, we have to accept a relatively large constant collector current. As we see from Fig. 201, a slope of the order of 100 mA/V can be obtained with an OC 71 type transistor, if the working point of the transistor is placed at $-I_C = 5$ mA. In the next chapter we shall show that the power gain is a maximum if the load resistance (or the load impedance) in the collector circuit is very close to the output impedance of the transistor.

Let us assume that we do in fact select the working point of an OC 71 transistor so that the constant collector current is 5 mA. The $-I_C = f(-V_{CE})$ characteristic of this transistor shows that its output impedance is approximately 10 kΩ. It is clear, however, that with a supply voltage of 10 V, for example, and a collector current of 5 mA, the load resistance must be much lower than 10 kΩ, in connection with the voltage drop taking place across it. Mutatus mutandi, for a load resistance R_L of 10 kΩ, the constant collector current must be made considerably lower than 5 mA, e.g. $-I_C = 0.5$ mA (see Fig. 202). At this value of the collector current, however, the slope of the OC 71 is much less than 100 mA/V.

This objection can be met by employing an inductive load (see Fig. 203). If the impedance of this load is $Z_L = 10\ \mathrm{k}\Omega$, and if its resistance can be neglected, the collector-emitter voltage $-V_{CE}$ will remain practically equal to the supply voltage $-V_{CC} = 10$ V. In this way it is possible to make full use of the large slope, but it should be remembered that low distortion in the amplifier stage may be just as desirable as a high current gain.

In the next chapter, it will be explained that the use of the conception of slope is only of value if the transistor is driven in a particular way, and that in most cases this gives rise to significant distortion. In addition, as already explained, a large collector current has to be accepted in order to obtain a large slope, at the cost of the efficiency of the stage.

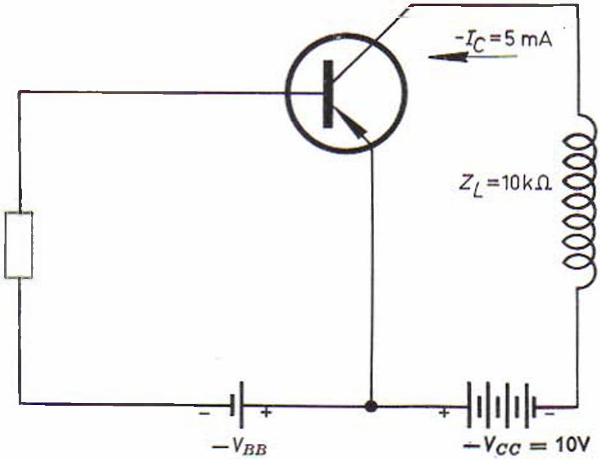

$-I_C = 5$ mA
$Z_L = 10\,k\Omega$
– +
$-V_{BB}$
+ –
$-V_{CC} = 10$V

Fig. 203

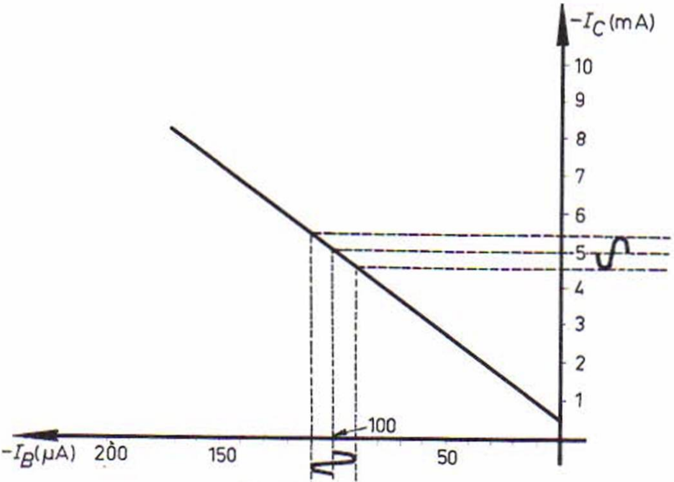


Fig. 204

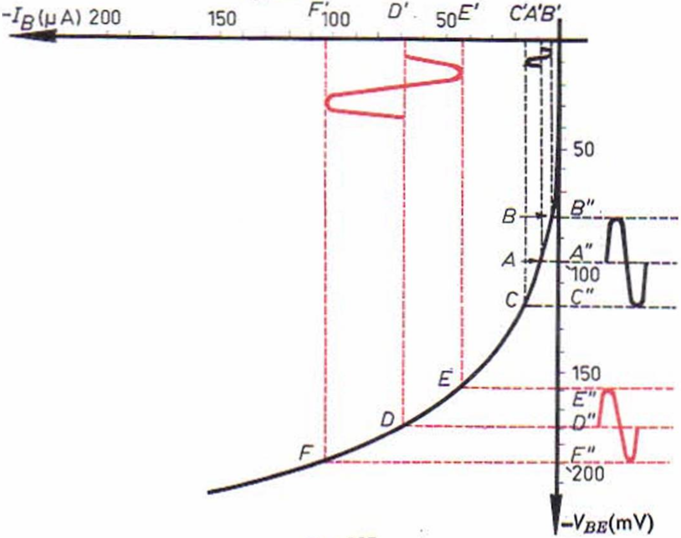


Fig. 205

The dynamic slope of a radio valve can be calculated from its equivalent circuit; this comprises an input circuit and an output circuit. By definition then, the slope equals the ratio of the alternating current in the output circuit, to the voltage across the input terminals. If we look at a transistor in the same way, with an output circuit and an input circuit, the collector current, i.e. the current in the output circuit, is found to depend on the base current, i.e. on the current flowing in the input circuit (see Fig. 204).

Variations in the base-emitter voltage are accompanied by variations in the base current, or in other words, variations of the input voltage are accompanied by variations in the driving current. Suppose that point A on the $-I_B = f(-V_{BE})$ characteristic is chosen as the working point, so that $-V_{BE} = 100$ mV, point A'' (Fig. 205). We shall also assume that this voltage varies by 20 mV on each side of the working point. Points B'' and C'' thus determined on the $-V_{BE}$ axis define points B and C on the characteristic. The variations of base current which are caused by these variations in the base-emitter voltage are given by the projections of points A, B, and C on the $-I_B$ axis (A', B' and C'). The amplitude of this current variation corresponds to approximately plus and minus 10 µA. If we had chosen the working point so that the constant base-emitter voltage was 180 mV, corresponding to point D on the characteristic, and to points D'' and D' on the $-V_{BE}$ axis and on the $-I_B$ axis respectively, the same variations in the base-emitter voltage (D'', E'', F''), would now correspond to base-current variations of approximately plus or minus 30 µA (D', E', F'). It is thus evident that the base-current variations which are caused by variations in the base-emitter voltage depend on the working point which is chosen, i.e. on the operating conditions of the transistor.

The effect described on the previous page only occurs in the base-emitter circuit of the transistor, that is, in the equivalent input circuit. For small-signal transistors this is the only factor which influences the slope. The $-I_C = f(-I_B)$ characteristic of this type of transistor can be regarded as practically straight, so that the relationship between collector-current variations and variations in the base-emitter voltage is determined exclusively by the relationship between base-current variations and variations in the base-emitter voltage. This is not true of transistors for higher powers. The $-I_C = f(-I_B)$ characteristic of these transistors is non-linear, so that the relationship between collector-current variations and variations of the base-emitter voltage also depends on the relationship between collector-current variations and base-current variations (see Fig. 206).

This effect is explained further in Fig. 207 in which the black curve represents the $-I_C = f(-I_B)$ characteristic of a higher–power transistor. Variations of the base current give rise to the collector-current variations which are printed black in the figure. If the $-I_C = f(-I_B)$ characteristic was linear (printed in red), the same variations of base current would result in greater variations of collector current.

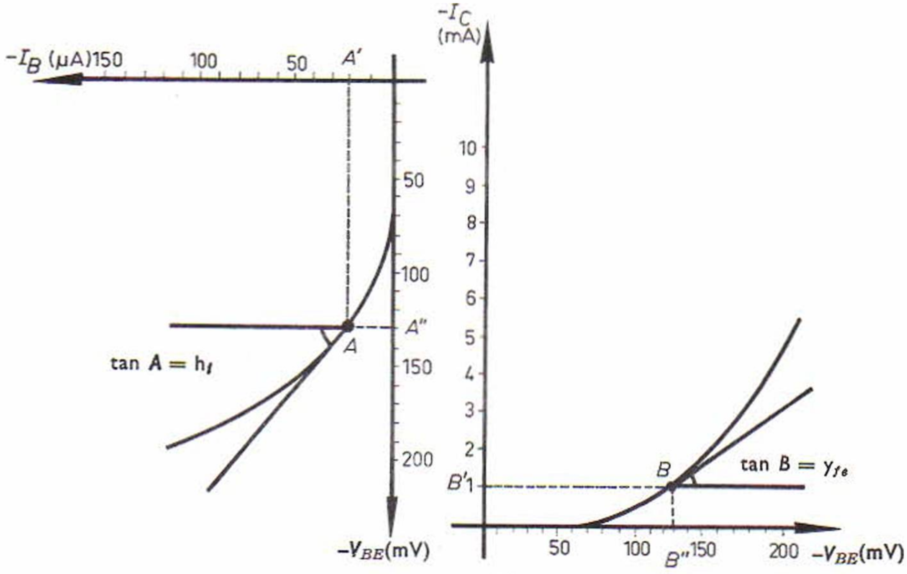


Fig. 206

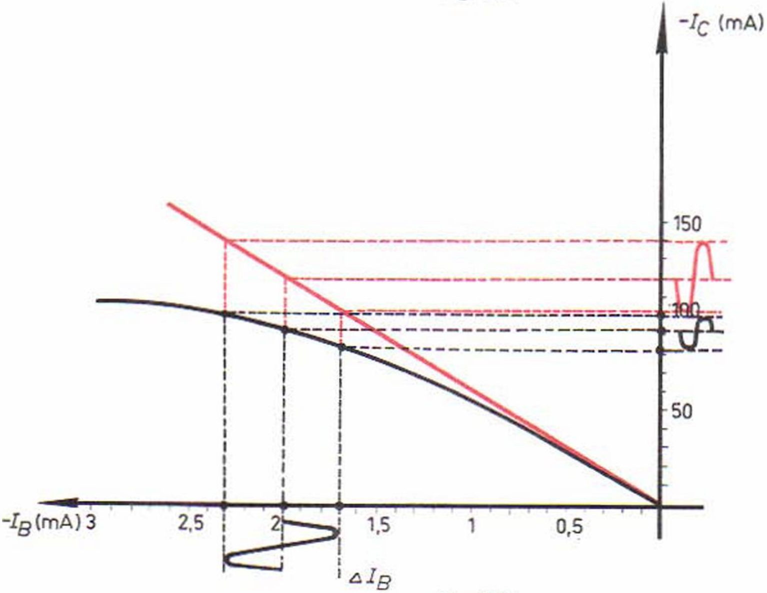
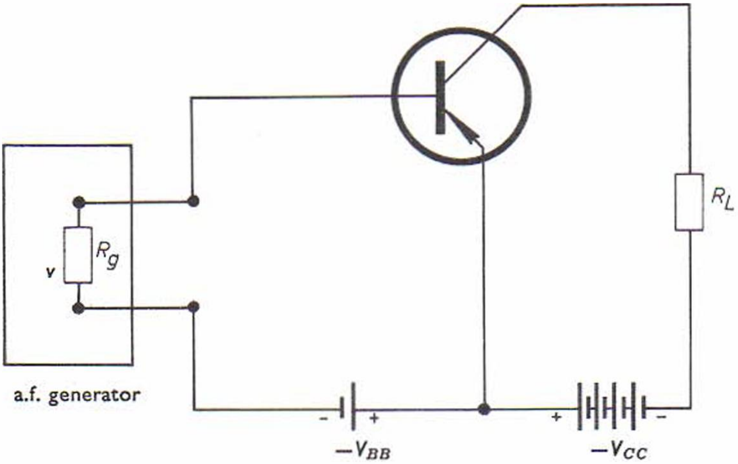
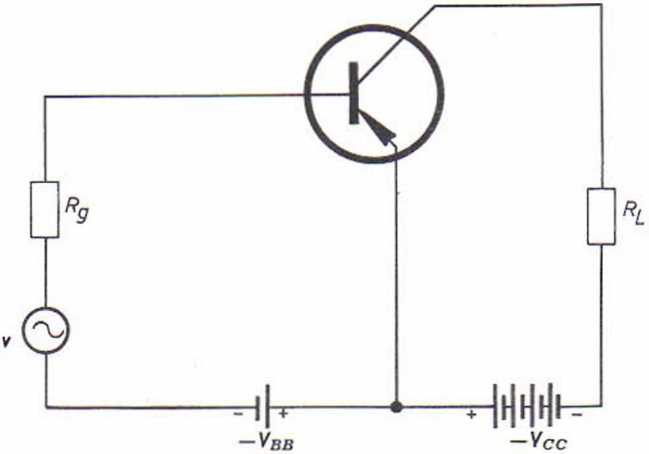


Fig. 207



v
R_g
a.f. generator
R_L
- +
$-V_{BB}$
+ -
$-V_{CC}$

Fig. 208



R_g
v
R_L
- +
$-V_{BB}$
+ -
$-V_{CC}$

Fig. 209

CHAPTER 22

Methods of driving small-signal transistors

The input impedance of the transistor is a factor which must be taken constantly into account in the design of an amplifier stage. This is because the method of driving a transistor is in the first place a question of matching. This matching affects the following two extremely important factors:

The power gain of the circuit;

The occurrence of non-linear distortion.

We have already explained on page 157 how the input load R_g can be represented in the characteristics of the transistor. We shall again assume that the base-emitter circuit is driven by a generator (see Fig. 208). This can be regarded as a voltage generator, the internal resistance of which is zero, and which is connected across the terminals of a circuit in which the generator resistance is connected in series with the input impedance h_i of the transistor (Fig. 209).

If the base-emitter voltage $-V_{BE}$ on the $-I_B = f(-V_{BE})$ characteristic of Fig. 150 is adjusted so that it corresponds to point A on the curve, the angle between the input load line and the horizontal will increase as the resistance of the generator increases.

We will now examine the following three cases separately:

Drive by a source with low internal resistance (voltage drive)
Drive by a source with high internal resistance (current drive)
Drive by a source whose internal resistance is of the same order of size as the input impedance of the transistor (mixed current-voltage drive).

The method of driving the transistor has a great effect on the gain and on the linear distortion of an amplifier stage.

22.1. Voltage drive

Fig. 210 represents a transistor with voltage drive. A generator having a very low internal resistance is connected to the base-emitter circuit of the transistor. Let us assume that this resistance is $R_g = 10\Omega$ and that the input impedance of the transistor $h_i = 1000\Omega$. Fig. 211 represents the equivalent circuit.

Now the input impedance of a transistor varies with the value of the base voltage, i.e. with its input voltage. If this is a pure sine wave, the input impedance will swing between two limits, which means that the base current will not be directly proportional to the applied voltage. In the case under discussion, the voltage supplied by the generator is applied directly between the base and the emitter of the transistor, so that we are dealing here with voltage drive.

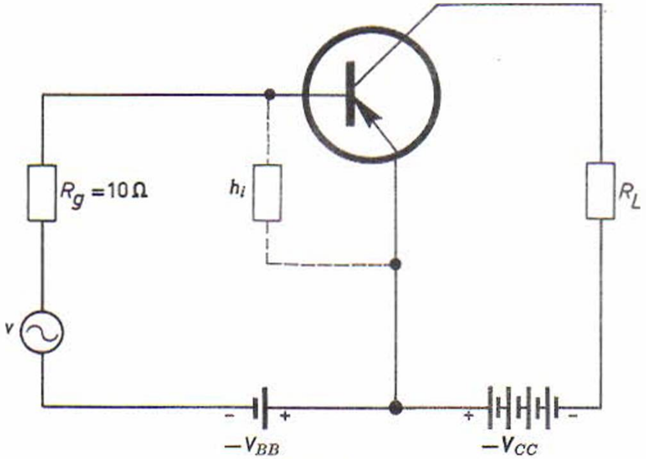

$R_g = 10\,\Omega$
h_i
R_L
v
−
+
$-V_{BB}$
+
−
$-V_{CC}$

Fig. 210

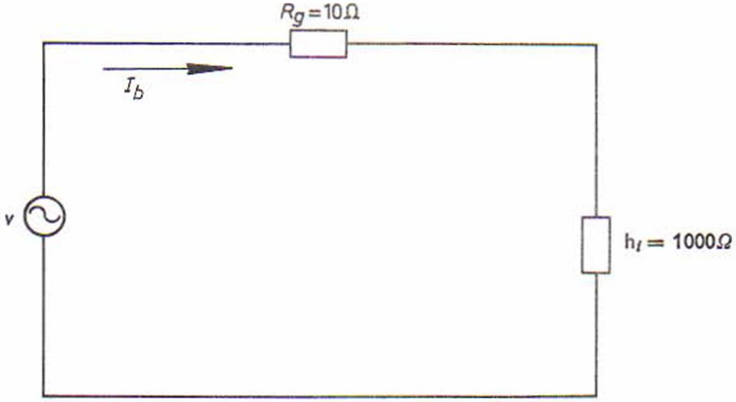

$R_g = 10\,\Omega$
I_b
v
$h_i = 1000\,\Omega$

Fig. 211

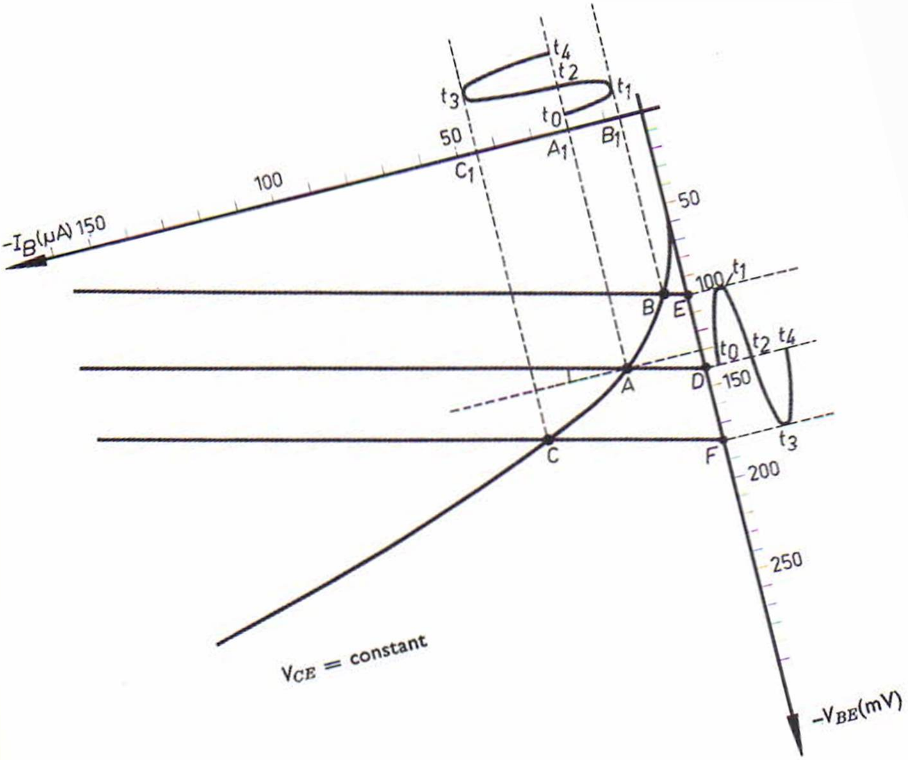

t4
t2
t1
t3
t0
50
A1
B1
C1
100
-IB(μA) 150
50
100
t1
B
E
t0
t2
t4
A
D
150
t3
C
F
200
250
VCE = constant
-VBE(mV)

Fig. 212

Let us examine this case more closely on the basis of the $-I_B = f(-V_{BE})$ characteristic of Fig. 212. The constant base-emitter voltage is 130 mV, which corresponds to a constant base current of 20 µA. These two values determine the point A on the characteristic.

Through this point, we draw a straight line making an angle A with the horizontal axis, such that tan A represents the input load R_g; this line determines point D on the $-V_{BE}$ axis. We must now make a distinction between two cases:

The transistor driven by signals of large amplitude (large signals).
The transistor driven by signals of small amplitude (small signals).

Large-signal drive

Let us suppose that the unloaded generator supplies a voltage of peak-to-peak value 80 mV. At instant t_0 the load line passes through A on the characteristic. From t_0 to t_1, the alternating voltage supplied by the generator will increase from zero to a maximum value, and the load line will move parallel to itself, until at instant t_1 it passes through E on the $-V_{BE}$ axis, and defines point B on the characteristic.

From t_1 to t_2, the generator voltage decreases again to zero, so that at instant t_2, the load line is once more in its original position. From t_2 to t_3 the generator voltage decreases from zero to a maximum negative value; the input load line now moves parallel to itself in the direction of the voltage variation, so that at instant t_3, it passes through F on the $-V_{BE}$ axis, and intersects the characteristic at point C. From t_3 to t_4 the generator voltage decreases from the maximum negative value to zero. At instant t_4 the load line is again in its original position.

The movement of the input load line as a function of the alternating voltage applied to the input thus defines points A, B and C on the $-I_B = f(-V_{BE})$ characteristic. The projections of these three points on the $-I_B$ axis (A_1, B_1 and C_1) give the corresponding variations of the base current. This shows that with a sinusoidal input voltage the corresponding input current is not sinusoidal.

Once we know the variations of the base current, it is possible to determine the corresponding variations of the collector current with the aid of the $-I_C = f(-I_B)$ characteristic. Suppose that A_1, B_1 and C_1 are the projections of points A, B and C on the $-I_B$ axis; these points determine points A_2, B_2 and C_2 on the $-I_C = f(-I_B)$ characteristic. The projections of the latter points on the $-I_C$ axis provide us with A_3, B_3 and C_3, which give the corresponding variations of collector current (Fig. 213).

We will now re-examine the $-I_C = f(-V_{BE})$ characteristic of the transistor (see Fig. 214). As we explained on page 201, this characteristic can be derived from the $-I_C = f(-I_B)$ and $-I_B = f(-V_{BE})$ characteristics. If we mark point A'' on the $-V_{BE}$ axis corresponding to $-V_{BE} = 130$ mV, this gives us point A on the characteristic and point A' on the $-I_C$ axis. This curve now enables us to read directly the variations of collector current produced by a sinusoidal base-emitter voltage.

From t_0 to t_1 the base-emitter voltage changes from A'' to B'', corresponding to point B on the characteristic and B' on the $-I_C$ axis. From t_1 to t_2 the base-emitter voltage changes from B'' to A'' corresponding to point A on the characteristic and point A' on the $-I_C$ axis. From t_2 to t_3 the base-emitter voltage changes from A'' to C'', corresponding to point C on the characteristic and point C' on the $-I_C$ axis. From t_3 to t_4 the base-emitter voltage changes from C'' to A'', again corresponding to point A on the characteristic and point A' on the $-I_C$ axis.

The variations of the collector current with respect to the base-emitter voltage can thus be determined directly with the aid of the $-I_C = f(-V_{BE})$ characteristic. For voltage drive, the slope of this characteristic is a very useful measure of the available gain. As has been shown, however, when this method of driving is applied to the amplification of large signals, it can lead to considerable distortion. However, this comment is not of general applicability, as we have assumed that the $-I_C = f(-I_B)$ characteristic is absolutely linear (with low output power). If the transistor operating conditions are selected so as to deliver greater power, the distortion will largely be cancelled by the curvature of this characteristic.

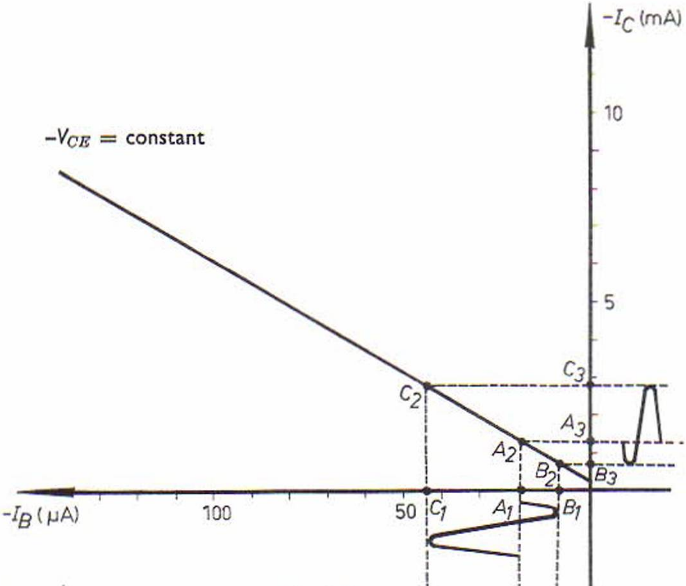


Fig. 213

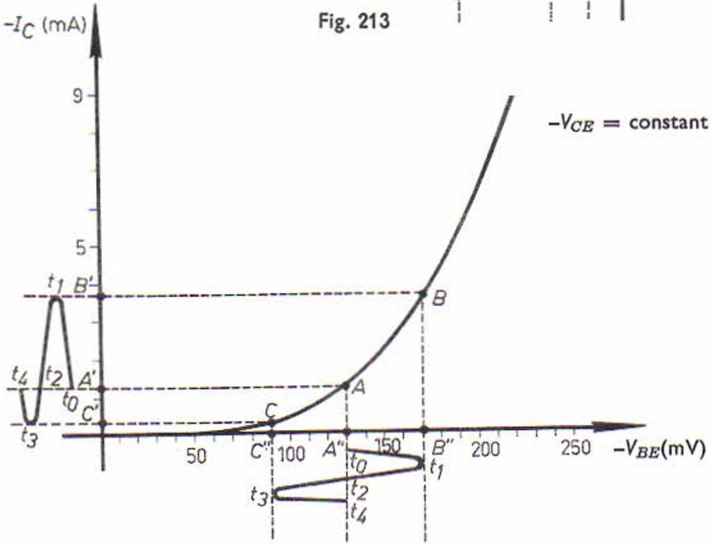


Fig. 214

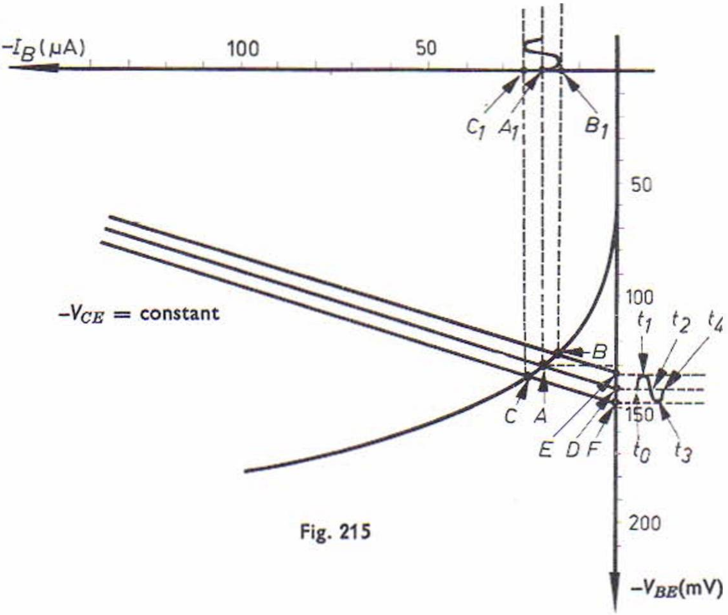


Fig. 215

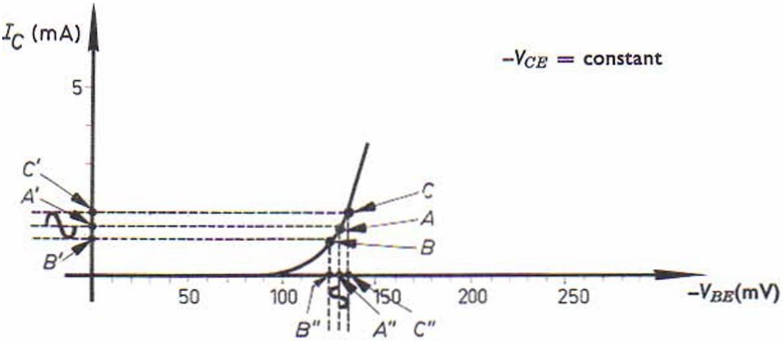


Fig. 216

Small-signal drive

Let us assume that the unloaded a.f. generator supplies a voltage of peak-to-peak value 10 mV (Fig. 215), and that the input load line cuts the $-I_B = f(-V_{BE})$ characteristic at instant t_0 at point A. As described in the previous case, the variations of the input alternating voltage will cause the load line to move to either side of this position.

The peak values of this voltage determine points E and F on the $-V_{BE}$ axis and points B and C on the characteristic. By projecting points A, B and C onto the $-I_B$ axis, we obtain points A_1, B_1 and C_1.

We now find that with a sinusoidal variation of the base-emitter current, the variation of the base current is also (practically) sinusoidal. This must be ascribed to the fact that if only a small portion of the curve is utilised, this portion can be regarded as a straight line.

As in the previous case, we can again make use of the $-I_C = f(-V_{BE})$ characteristic (Fig. 216). If we mark on this curve the point $-V_{BE} = 130$ mV (point A''), this corresponds to point A on the characteristic and point A' on the $-I_C$ axis.

The variations of the base-emitter voltage are represented by the peak values of the input voltage, that is by points B'' and C'' on the $-V_{BE}$ axis, corresponding to points B and C on the characteristic and points B' and C' on the $-I_C$ axis. As for large input signals, the gain of a voltage-driven transistor for small input signals is given by the slope of the $-I_C = f(-V_{BE})$ characteristic.

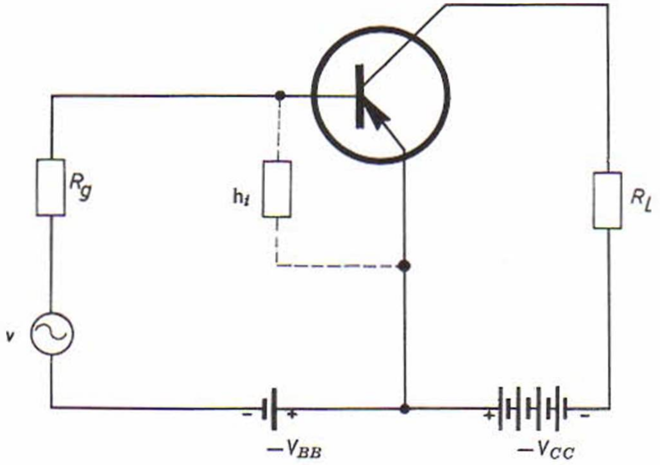
22.2. Current drive

Fig. 217 represents the transistor with current drive. The a.f. generator in the input circuit now has a very high impedance, e.g. $R_g = 1\ M\Omega$. The equivalent circuit is given in Fig. 218. In this case the current I_B can be taken as equal to $-V_{BB}/R_g$, as the input impedance h_i of the transistor can now be neglected in relation to R_g. This means that the current in the input circuit is determined solely by the generator impedance, and that variations of the input impedance of the transistor as a function of the base-emitter voltage can have no effect on the current. If the input voltage V is sinuoidal, the base current I_b will also be sinusoidal.

The base-emitter voltage V_{be} is equal to the product of the base current and the input impedance of the transistor;

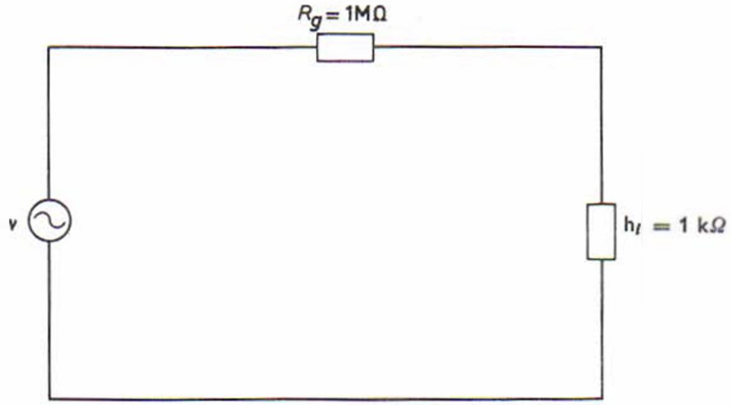
$$V_{be} = I_b . h_i.$$

Since the input impedance of the transistor is subject to variation, the base-emitter voltage resulting when the transistor is driven by a sinusoidal voltage will not be sinusoidal. We refer to this as current drive of the transistor.



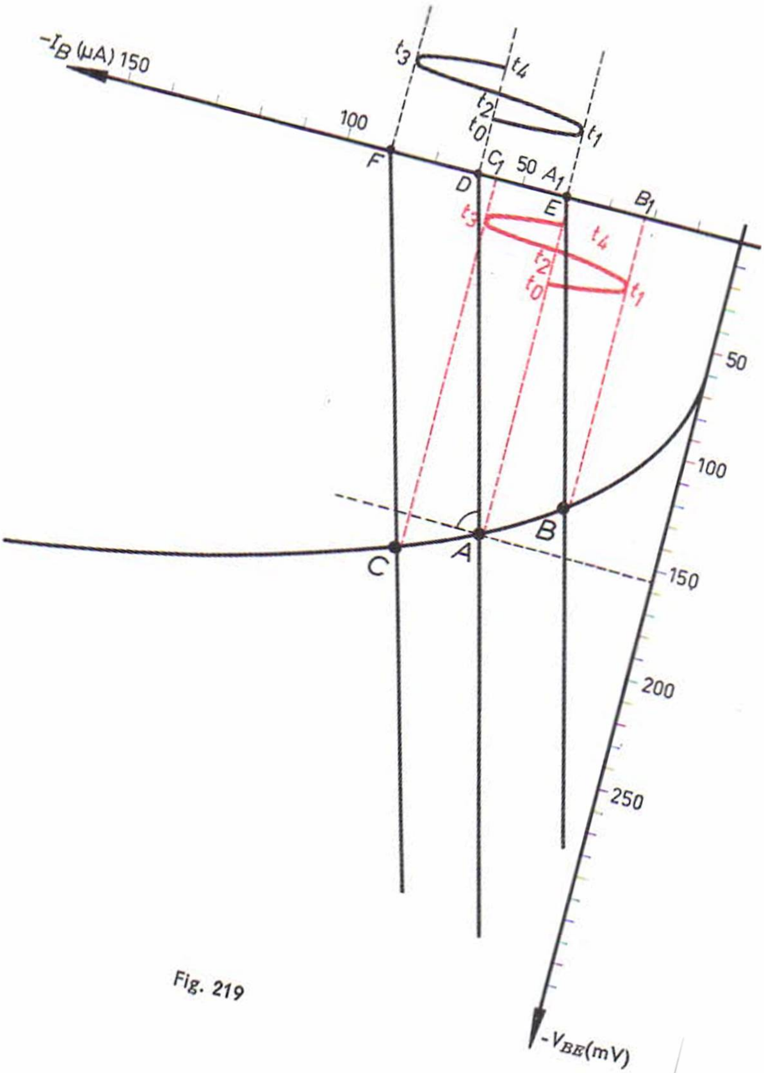
R_g
h_i
R_L
v
$-V_{BB}$
$-V_{CC}$

Fig. 217



$R_g = 1M\Omega$
v
$h_i = 1\ k\Omega$

Fig. 218



$-I_B$ (μA) 150
100
50
t_3
t_4
t_2
t_0
t_1
F
D
C_1
A_1
E
B_1
t_3
t_4
t_2
t_0
t_1
50
100
150
200
250
A
B
C
$-V_{BE}$(mV)

Fig. 219

Let us examine the $-I_B = f(-V_{BE})$ characteristic of the transistor (Fig. 219). Assume that the working point is chosen so that $-V_{BE} = 154$ mV, corresponding to a base current $-I_B = 40$ µA. The working point is represented by point A on the characteristic. The input load line through this point will be very steep because tan $A = R_g$ is very large, and angle A thus approaches 90°. This load line determines point D on the $-I_B$ axis. Once again, we will restrict ourselves for the time being to considering the transistor with operating conditions adjusted to give a small output power, so that the $-I_C = f(-I_B)$ characteristic is practically a straight line. Let us assume that the current supplied by the generator has a peak-to-peak value of 40 µA so that the base current varies between 40 and 80 µA. From t_0 to t_1 the alternating current increases from zero to a maximum positive value, and the input load line moves parallel to itself from point D to E; at instant t_1, this line intersects the characteristic at point D. From t_1 to t_2, the alternating current decreases from the maximum positive value to zero, and the load line takes up its original position once more. From t_2 to t_3 the alternating current changes from zero to a negative maximum, and the load line moves parallel to itself towards the other side, so that at instant t_3 it intersects the characteristic at C and defines a point F on the $-I_B$ axis. From t_3 to t_4 the alternating current again returns to zero, and the load line once more takes up its original position (Point E on the $-I_B$ axis and point A on the characteristic).

By projecting points A, D and C onto the $-I_B$ axis, we obtain points A_1, B_1 and C_1 representing the variations of the transistor base current (printed in red). We see that these variations are practically sinusoidal, corresponding to the sinusoidal input voltage derived from the generator. This shows that current drive offers the possibility of amplifying relatively large signals with much less distortion than is possible with voltage drive.

22.3. The effect of the generator impedance on the internal feedback of the transistor

Voltage drive

With voltage drive, the generator has a low impedance (much lower than the input impedance of the transistor). Now the equivalent circuit for the internal feedback contains the impedance between the base and the collector of the transistor, with any external circuit elements connected in parallel, and also the impedance between the base and the emitter with any external circuit elements that may be connected in parallel. (see Fig. 220).

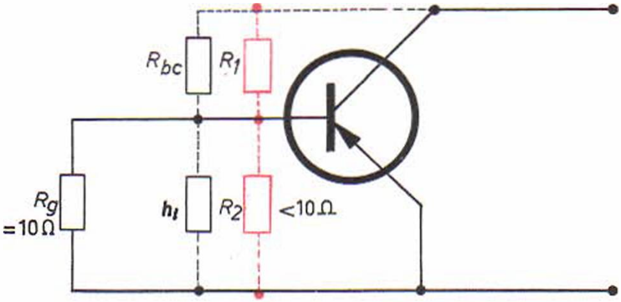
If we represent the equivalent resistance of the base-collector circuit by R_1 and the equivalent resistance of the base-emitter circuit by R_2, we have:

$$R_1 = h_o \qquad \text{and } R_2 = h_i R_g/(h_i + R_g).$$

Now with voltage drive R_g is extremely small, or in other words the equivalent resistance of the input circuit becomes very low when viewed from the output circuit. Since the feedback is given by the quotient R_2/R_1, both it and R_2 will take on a very low value. This means that with voltage drive the internal feedback of the transistor can be neglected.

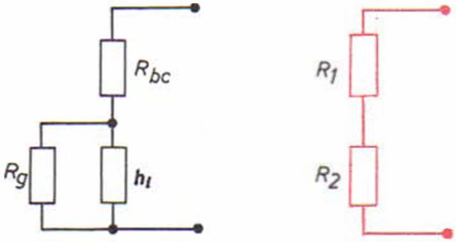
Current drive

With current drive the generator has a high impedance which far exceeds the input impedance of the transistor. In this case also, the internal feedback is of course determined by the quotient R_2/R_1. Here, however, R_2 is practically equal to the input impedance of the transistor, since the impedance connected in parallel is relatively large (see Fig. 222). With current drive the internal feedback of a transistor thus approaches the maximum value h_{re}, as described on page 181.



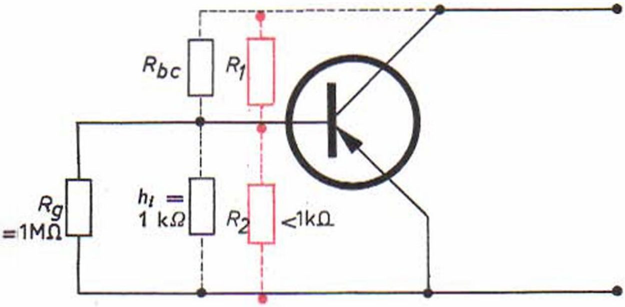
R_{bc}
R_1
R_g
$=10\,\Omega$
h_i
R_2
$<10\,\Omega$

Fig. 220



R_{bc}
R_g
h_i
R_1
R_2

Fig. 221



R_{bc}
R_1
R_g
$=1\,M\Omega$
$h_i =$
$1\,k\Omega$
R_2
$<1\,k\Omega$

Fig. 222

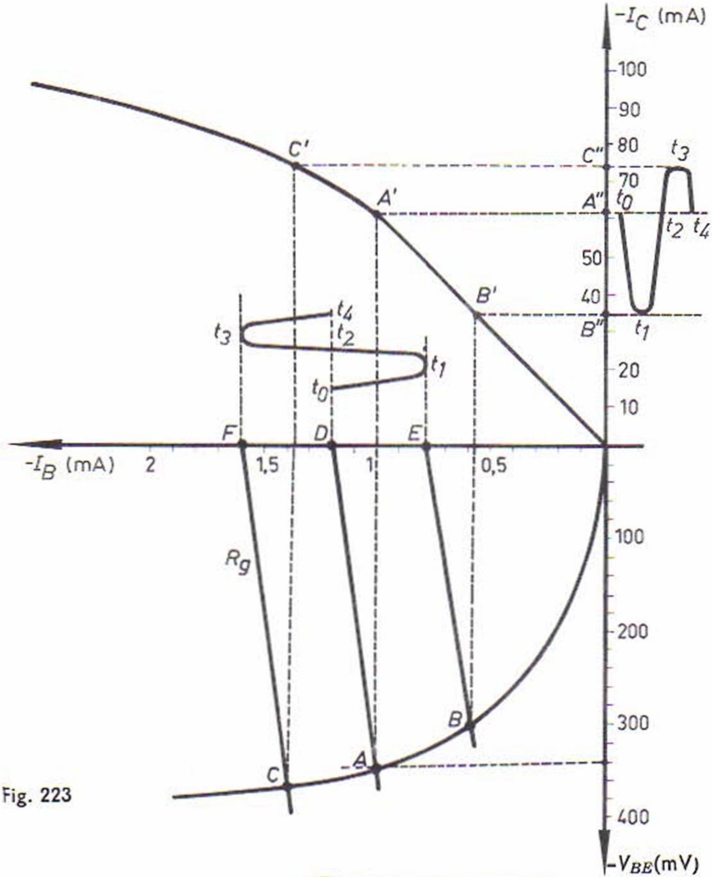
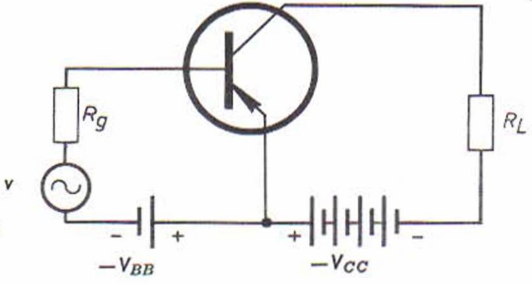


Fig. 223



R_g
R_L
v
–
+
+
–
$-V_{BB}$
$-V_{CC}$

Fig. 224

CHAPTER 23

Methods of driving power transistors

If a transistor is to supply a large output power, the curved part of the $-I_C = f(I_B)$ characteristic (Fig. 223) will be employed.

23.1. Current drive

Let us assume that a transistor is adjusted to a base-emitter voltage of 340 mV corresponding to a constant base current $-I_B$ of 1 mA and a constant collector current $-I_C$ of 62 mA (point A on the $-I_B = f(-V_{BE})$ characteristic). We will first consider the case in which the transistor is driven by a generator with a very high resistance (current drive). The input load line is then represented by a very steep straight line. The load line through point A defines point D on the $-I_B$ axis.

Here too, the variations of current in relation to point D on the input circuit will affect the position of the input load line. From t_0 to t_1 the load line moves from point D to point E, cutting the $-I_B = f(-V_{BE})$ characteristic at point B. From t_2 to t_3 the load line moves from point D to point F, so that the $-I_B = f(-V_{BE})$ characteristic is cut at point D. From t_3 to t_4 the load line returns to its original position once more.

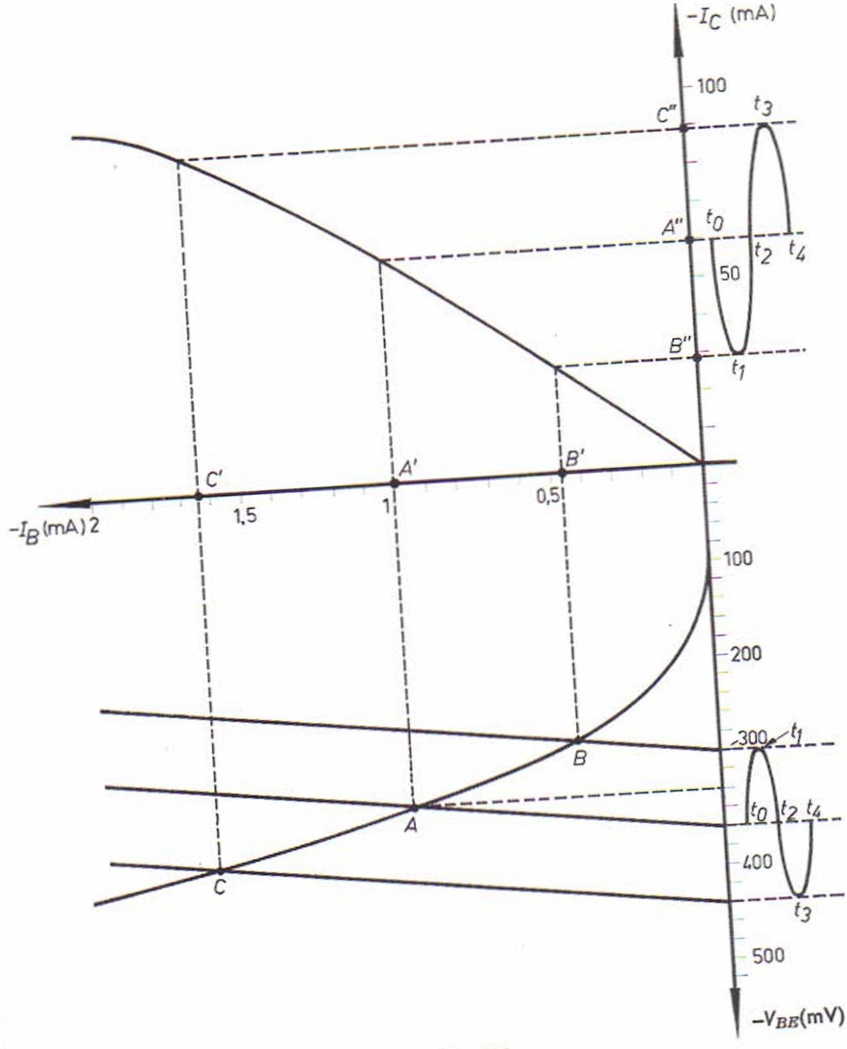
The movement of this input load line as a function of the variations of current in the input circuit defines points A', B' and C' on the $-I_B = f(-I_C)$ characteristic, and the projection of these points onto the $-I_C$ axis (A'', B'' and C'') show that the sinusoidal base current results in a non-sinusoidal collector current. This means that when a transistor is being used to give a high power output, the curvature of the $-I_B = f(-I_C)$ characteristic does not guarantee distortion-free amplication of large signals with current drive. We will now see whether voltage drive offers an improvement in this respect.

23.2. Voltage drive

The distortion which may occur when a transistor is current driven, as demonstrated on the previous page, must be ascribed to the curvature of the $-I_C = f(-I_B)$ characteristic. In this respect, voltage drive is frequently to be preferred.

Fig. 225 shows the characteristics of the OC72 transistor. From the figure we can see that the curvature of the $-I_C = f(-I_B)$ characteristic is partially compensated by the curvature of the $-I_B = f(-V_{BE})$ characteristic. Suppose that the base-emitter voltage in the absence of a signal is 320 mV. The constant base current then equals 1 mA and the constant collector current is 60 mA. Point *A* represents the working point on the $-I_B = f(-V_{BE})$ characteristic. As the impedance of the generator which is driving the transistor is now very low, the input load line through this point will be almost horizontal. Let the peak-to-peak value of the input alternating voltage be 160 mV. As a result the load line will move symmetrically with respect to its original position.

The extreme positions define points *B* and *C* on the $-I_B = f(-V_{BE})$ characteristic. The projections of these points onto the $-I_B$ axis give the variation of the base current caused by variations of the base-emitter voltage: with these, we can determine the corresponding variations of the collector current. These are found to agree much better with the variations of the input voltage than they did in the previous case, which must be ascribed to the fact that the curvatures of the characteristics partly compensate each other. In this case, therefore, voltage drive proves to be preferable to current drive. The ideal method of driving a transistor would thus be to pass from current drive to voltage drive when the input signal exceeds a certain value.



$-I_C$ (mA)
100
C''
t_3
t_0
A''
t_2
t_4
50
B''
t_1
C'
A'
B'
$-I_B$(mA) 2
1,5
1
0,5
100
200
300
t_1
B
A
t_0
t_2
t_4
400
C
t_3
500
$-V_{BE}$(mV)

Fig. 225

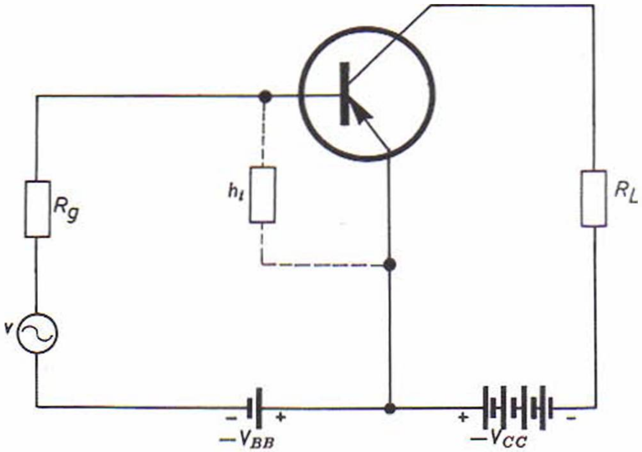

R_g
h_i
R_L
v
$-V_{BB}$
$-V_{CC}$
− +
+ −

Fig. 226

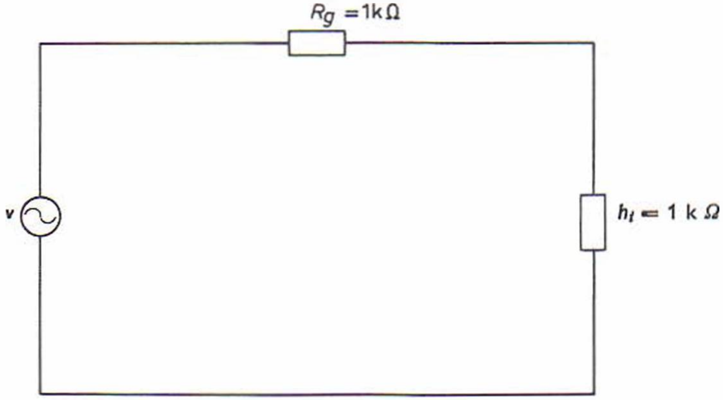

$R_g = 1\text{k}\Omega$
v
$h_i = 1\ \text{k}\ \Omega$

Fig. 227

CHAPTER 24

Matched drive

We refer to matched drive if the generator impedance is equal or practically equal to the input impedance of the transistor, or in other words if $R_g \approx \mathrm{h}_i$. We will explain this in more detail with the aid of Fig. 226.

Suppose that the generator impedance and the input impedance of the transistor both have a value of 1 kΩ; in this case we have the equivalent circuit of Fig. 227. The current flowing in this circuit is determined by the generator voltage and by the total circuit impedance, consisting of the generator impedance and the transistor impedance.
This current is therefore:

$$i = v/(R_g + \mathrm{h}_i).$$

Since R_g is equal to h_i the latter impedance cannot now be neglected, and a sinusoidal generator voltage of large amplitude will cause non-sinusoidal variations of the current. In its turn, the input voltage v_{be} will depend on the input impedance of the transistor and, with a sinusoidal input current of great amplitude, will deviate from the sinusoidal to the same extent. We are thus dealing with both current drive and voltage drive, and it will be found that if undistorted amplification is required, this method is only suitable for small amplitudes. Let us consider the operation of a transistor with matched drive for the following cases:

1) For small signals.
2) For large signals.

24.1. Small signals

Fig. 228 is an example of a circuit with matched drive. The transistor is driven by a generator having an impedance R_g of 1 kΩ. Its $-I_C = f(-I_B)$ and $-I_B = f(-V_{BE})$ characteristics are represented in Fig. 229. Let us assume that, in the absence of a signal, the base-emitter voltage $-V_{BE} =$ 130 mV, corresponding to a base current $-I_B = 20$ µA. The transistor is thus adjusted to point *A* on the $-I_B = f(-V_{BE})$ characteristic. Through this point we draw the input load line, so that the tangent *A* corresponds to the value of R_g. This line determines point *D* on the $-V_{BE}$ axis and point *G* on the $-I_B$ axis.

We can now look at the variations of driving voltage or driving current, produced by the generator. Suppose that these voltage variations have an amplitude of 20 mV; this will mean that the input load line moves so as to define points *E* and *F* on the $-V_{BE}$ axis, points *D* and *C* on the characteristic, and points *H* and *J* on the $-I_B$ axis. The last-mentioned points correspond to the amplitudes of the generator current with short-circuited output.

The projections of points *A*, *B* and *C* on the $-I_B$ axis determine the corresponding variations of the base current (*A'*, *B'* and *C'*); from these points it is possible to determine the variations of the collector current (*A"*, *B"* and *C"*) with the aid of the $-I_C = f(-I_B)$ characteristic. This graphical representation shows that the variations of collector current are almost sinusoidal, but this is true only for signals of small amplitude.

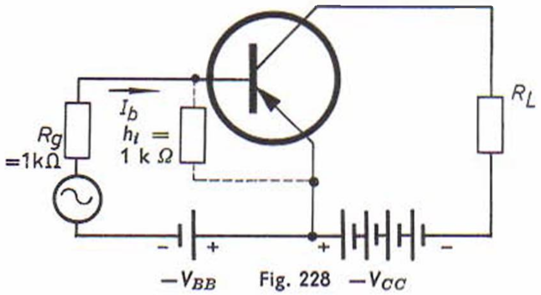

I_b
R_g =1kΩ
h_i = 1 kΩ
R_L
–
+
+
–
$-V_{BB}$
$-V_{CC}$

Fig. 228

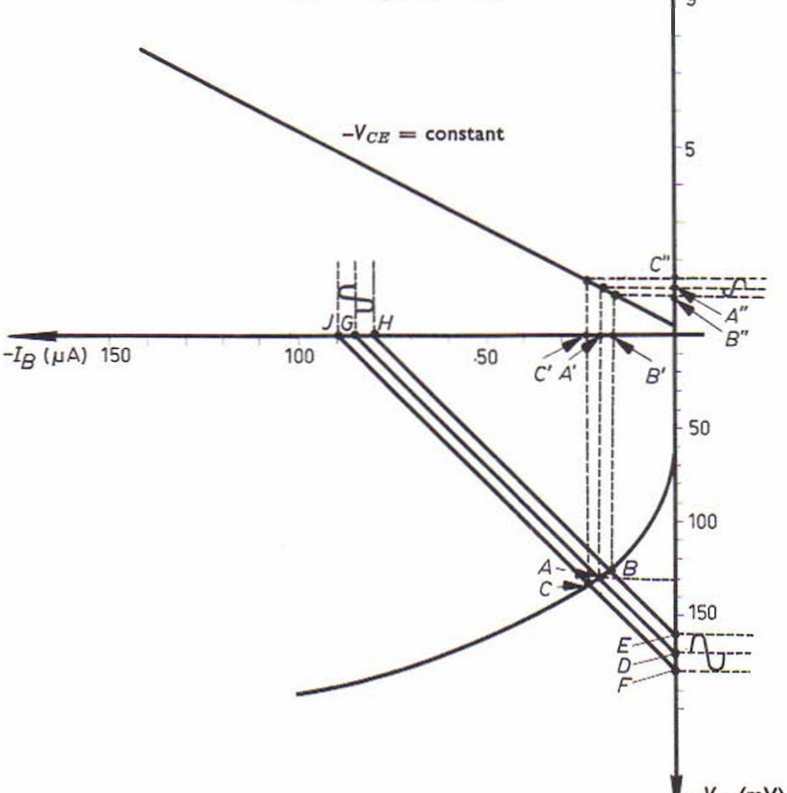


Fig. 229

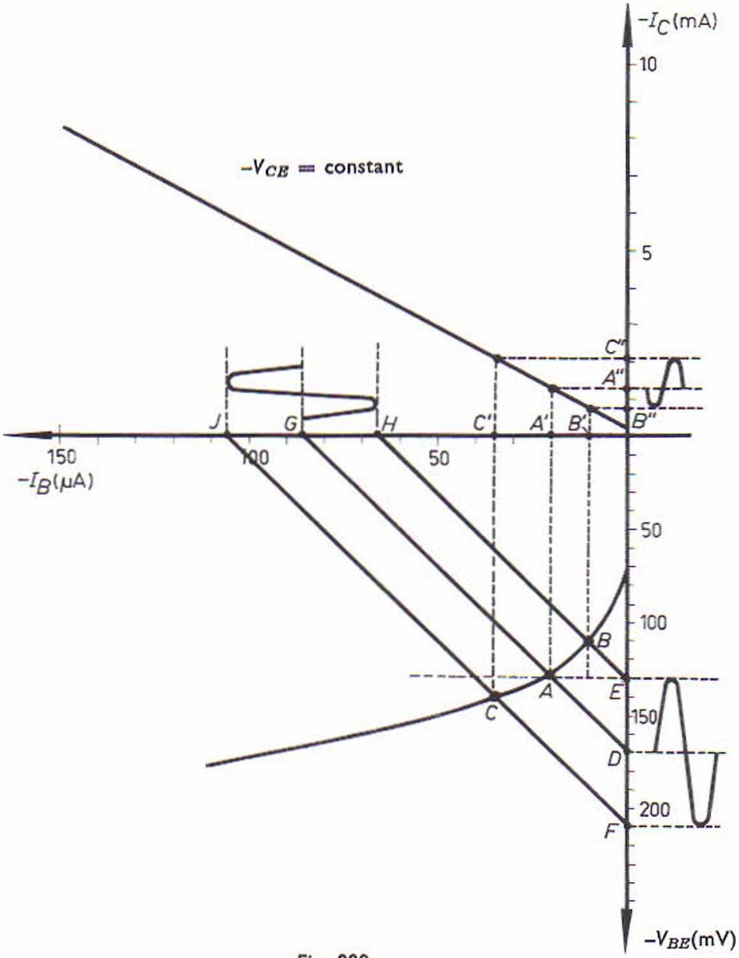

-I_C(mA)
10
5
-V_{CE} = constant
C''
A''
B''
J
G
H
C'
A'
B'
150
100
50
-I_B(μA)
50
100
B
A
E
C
150
D
200
F
-V_{BE}(mV)

Fig. 230

24.2. Large signals

In this case too, we will start from the circuit of Fig. 228. The $-I_B = f(-V_{BE})$ and $-I_C = f(-I_B)$ characteristics of the transistor are re-drawn in Fig. 230. Once again we assume that in the steady state $-V_{BE} = 130$ mV, corresponding to a base current of $-I_B = 20$ µA. Point A on the characteristic again represents the working point of the transistor in the absence of a signal. Once more we draw the load line through point A, so that the tangent of angle A corresponds to the value of R_g. This gives us point D on the $-V_{BE}$ axis and point G on the $-I_C$ axis.

Suppose that the control voltage supplied by the generator has a peak-to-peak value of 80 mV. This will mean that the input load line moves so as to determine points E and F on the $-V_{BE}$ axis, points B and C on the characteristic and points H and J on the $-I_B$ axis. The projections of points A, B and C onto the $-I_B$ axis give the variations of the base current. (Points H, G and J again represent the current variations of the generator with short-circuited output).

It will be seen that the variations of the collector current caused by the variation of base current are far from sinusoidal. Matched drive is thus accompanied by the same disadvantages as voltage drive, i.e. serious distortion occurs with large signals. This method of driving therefore appears to be subject to the same limitations as the two other methods of driving; in every case where current drive or voltage drive as the case may be cannot be employed, matched driving will be equally unsuitable. Consequently, the latter can only be recommended for amplifying small signals, when the distortion must remain slight.

CHAPTER 25

Distortion and power amplification

In the previous chapter we have discussed the different methods of driving a transistor and the effect that these have on the linearity of the gain. This gain is closely related to the following two factors:

The lowest possible distortion.
The maximum gain.

25.1. Minimum distortion

Fig. 231 shows the distortion as a function of the generator impedance when the transistor is operated in a linear portion of the $-I_C = f(-I_B)$ characteristic (small output power transistor). The same relationship is represented in Fig. 232, when the transistor is operating in a curved portion of the $-I_C = f(-I_B)$ characteristic (large output power transistor).
The great objection to voltage drive is that with signals of medium or large amplitude, high distortion occurs if the transistor is operated in the straight portion of the $-I_C = f(-I_B)$ characteristic (range *AB*, Fig. 231). Consequently, operating conditions of this nature can only be recommended for the amplification of small signals. If the transistor is operated in the curved portion of the $-I_C = f(-I_B)$ characteristic, voltage drive can only be recommended for amplifying very large signals (range *AB*, Fig. 232).
Current drive offers the possibility of linear amplification, independent of the signal amplitude, provided that the portion of the $-I_C = f(-I_B)$ characteristic which is employed is perfectly straight (range *CD*, Fig. 231). If this portion of the characteristic is curved (large output power), serious distortion will occur (region *CD*, Fig. 232). On the other hand, matched drive always leads to distortion with large signals, independent of the shape of the $-I_C = f(-I_B)$ characteristic of the transistor (range *BC*, Figs. 231 and 232).

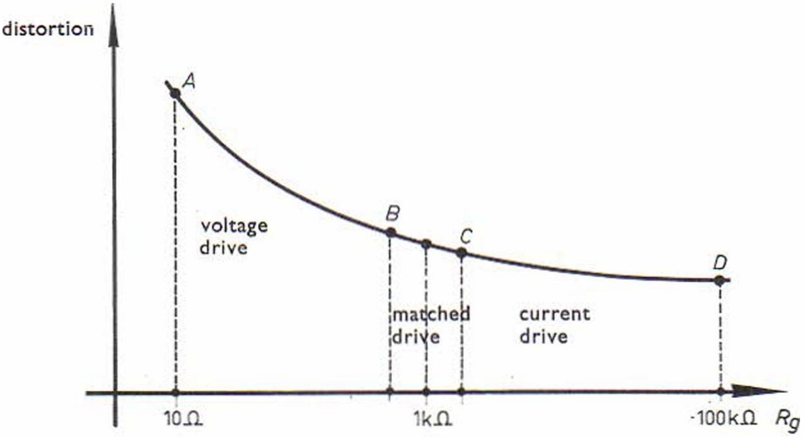


Fig. 231

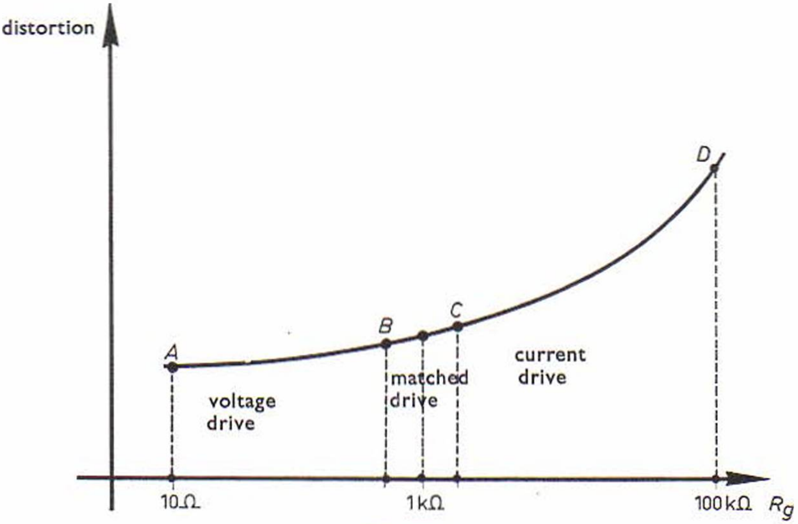


Fig. 232

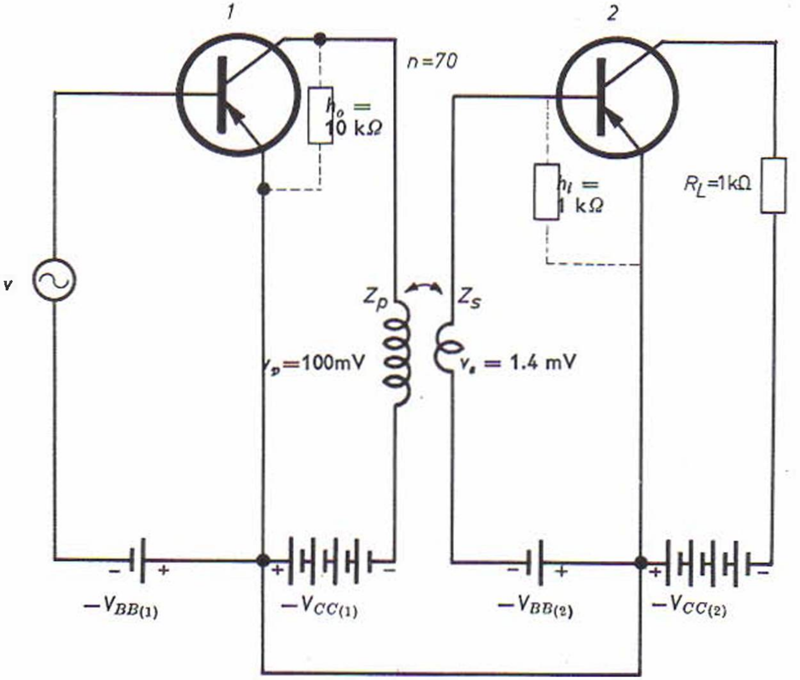

1
2
n=70
h_o = 10 kΩ
h_i = 1 kΩ
R_L=1kΩ
v
Z_p
Z_s
v_p=100mV
v_s = 1.4 mV
- +
+ -
$-V_{BB(1)}$
$-V_{CC(1)}$
$-V_{BB(2)}$
$-V_{CC(2)}$

Fig. 233

25.2. Power amplification

The object of the following paragraphs is to investigate the effect of the various methods of driving the transistor on the power gain of a transistor amplifier. For amplifying small signals, each one of the three methods of driving can be applied, so that it is useful to investigate which of these methods gives the largest power gain.

Voltage drive

In the circuit of Fig. 233 transistors 1 and 2 are coupled by means of a transformer, which we shall assume to have a turns ratio of 70 : 1. We shall also assume that the output impedance h_o of transistor 1 has a value of 10 kΩ. Transformed to the secondary, this gives the generator impedance of transistor 2 as:

$$Z_s = \mathrm{h}_o/n^2 = 10 \times 10^3/70^2 \simeq 2\Omega.$$

As the generator impedance of transistor 2 is very much lower than its input impedance h_i, the amplifier stage that we are considering is voltage driven. We will define the "power gain" of the stage as the quotient of the available power across the load resistance R_1 (of 1 kΩ) of transistor 2, and the maximum power which transistor 1 can supply.

Power available at the output

The power available at the output is equal to the product of the load resistance R_L and the square of the collector current i_c^2, that is:

$$P_{o2} = R_L\, i_c^2 = 1000\, i_c^2.$$

The value of i_c can be determined as follows. Suppose that the available voltage across the primary of the transformer i.e. the input voltage of this stage, is equal to

$$v_p = v_i = 0.1 \text{ V}.$$

so that the secondary voltage v_s is:

$$v_s = v_p/70 = 1.4 \text{ mV}.$$

The current flowing through the input circuit of transistor 2 is equal to the quotient of the voltage v_s across the input impedance of this transistor, and the value of this impedance h_i:

$$i_b = v_s/\mathrm{h}_i = 1.4 \times 10^{-3}/10^3 = 1.4 \times 10^{-6} \text{ A} = 1.4\ \mu\text{A}.$$

Let us assume that the current gain of transistor 2 is 50; the collector current is then:

$$i_c = 50 \times 1.4 = 70\ \mu\text{A}.$$

The available output power of this amplifier stage is therefore:

$$P_o{}^2 = R_L i_c{}^2 = 10^3.\ (70 \times 10^{-6})^2 = 49 \times 10^{-7}\ \text{W} \simeq 5\ \mu\text{W}.$$

Maximum output power of transistor 1

The maximum output power that transistor 1 can supply is obtained when the load impedance Z_L of this transistor is equal to its output impedance h_o (see Fig. 234a), and in that case is:

$$P_{o1\text{max}} = \tfrac{1}{2}E^2/(\text{h}_o + Z_L) = \tfrac{1}{4}E^2/\text{h}_o$$

where E is the output voltage of the unloaded transistor.

Now in the case under consideration, the load impedance of the transistor is equal to the input impedance of transistor 2, transformed to the primary:

$$Z_p = Z_L = n^2\ \text{h}_i = 70^2 \times 10^3 = 49 \times 10^5 \Omega = 4.9\ \text{M}\Omega.$$

The output impedance h_o of the transistor may be neglected in comparison with this impedance, so we may assume that the output voltage E in the unloaded condition is equal to the value of 0.1 V assumed for v_p, so that:

$$P_{o1\text{max}} = \tfrac{1}{4}\ E^2/\text{h}_o = \tfrac{1}{4}.v_i{}^2/\text{h}_o = \tfrac{1}{4}.10^{-2}/10^4 = 0.25 \times 10^{-6}\ \text{W} = 0.25\ \mu\text{W}.$$

To prevent any ambiguity it should be pointed out that this power is appreciably greater than the power which is actually delivered to the load. This is because the latter is equal to the quotient of the square of the voltage across the load impedance, and this impedance, which is 4.9 MΩ in the present case (see Fig. 234b):

$$P_{o1} = v_i{}^2/Z_L = 10^{-2}/(49 \times 10^5) \simeq 2 \times 10^{-9}\ \text{W} = 0.002\ \mu\text{W}.$$

Power gain

The power gain, that is, the ratio of the power available at the output of transistor 2 to the maximum power that transistor 1 can supply, is consequently,

$$G_P = P_{o2}/P_{o1\text{max}} = 5/0.25 = 20.$$

in the case under consideration (with voltage drive).

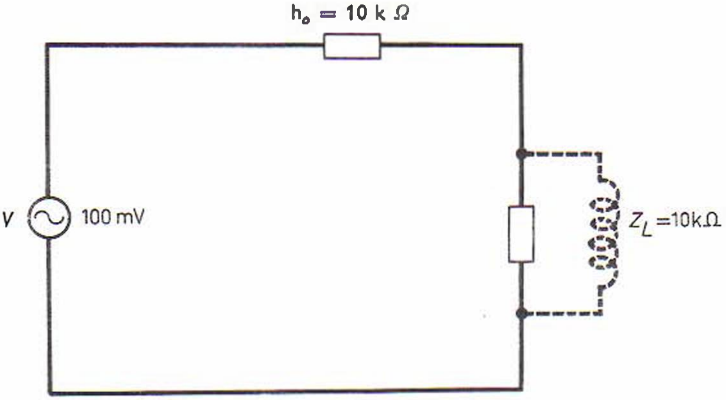

$h_o = 10\ k\ \Omega$
v
100 mV
$Z_L = 10 k\Omega$

Fig. 234a

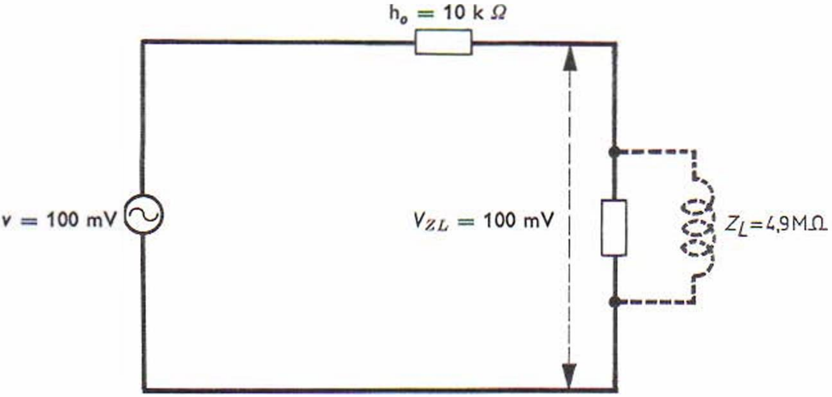

$h_o = 10\ k\ \Omega$
$v = 100$ mV
$V_{ZL} = 100$ mV
$Z_L = 4{,}9 M\Omega$

Fig. 234b

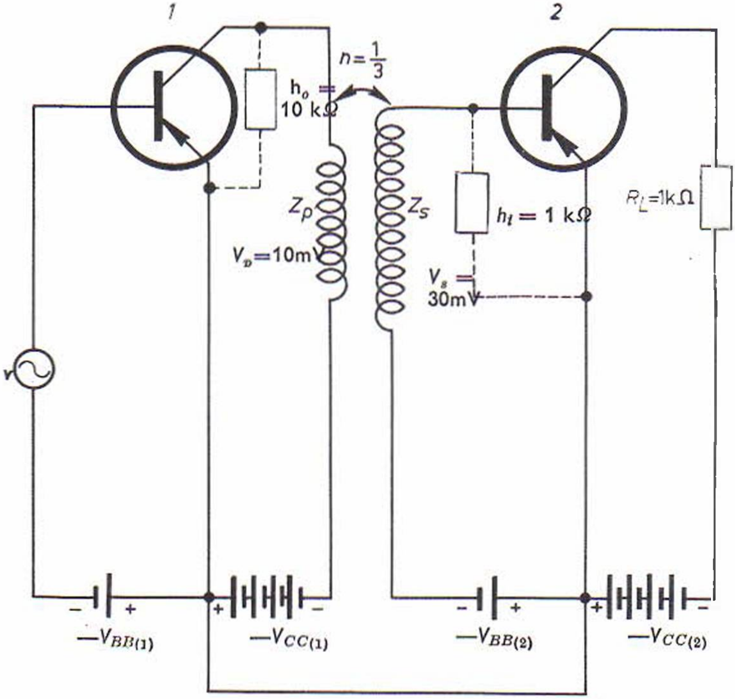

1
2
n = 1/3
h_o = 10 kΩ
Z_P
Z_S
V_P = 10mV
h_i = 1 kΩ
V_S = 30mV
R_L = 1kΩ
v
– +
—V_BB(1)
+ –
—V_CC(1)
– +
—V_BB(2)
+ –
—V_CC(2)

Fig. 235

Current drive

In the circuit of Fig. 235, transistors 1 and 2 are coupled together by means of a transformer having a turns ratio of 1 : 3. The output impedance h_o of transistor 1 is again taken as equal to 10 kΩ, so that the output impedance of transistor 2, transformed to the secondary, is:

$$Z_s = h_o/n^2 = 10^4/(1/3)^2 = 90\,000\Omega = 90\text{ k}\Omega.$$

In this case, therefore, the generator impedance is much greater than the input impedance of transistor 2, so that the transistor is current driven. The power gain will now be calculated in the same manner as in the previous example.

Power available at the output

In this case also, the power available at the output is

$$P_{o2} = 3_L\, i_c^{\,2} = 1000\, i_c^{\,2}$$

in which, however, i_c has a different value. This depends on the base current, which in its turn is determined by the voltage across the secondary of the transformer and by the input impedance of the transistor. Suppose that the primary voltage v_p = 10 mV, so that the secondary voltage is 3 × 10 = 30 mV. The current flowing through the input circuit of transistor 2 is then:

$$i_b = v_s/h_i = 3 \times 10^{-2}/10^3 = 3 \times 10^{-5}\text{ A} = 30\ \mu\text{A}.$$

With a current gain of 50 the collector current will equal:

$$i_c = 50 \times 30 = 1500\ \mu\text{A} = 15 \times 10^{-4}\text{ A}.$$

The available power is thus:

$$P_{o2} = R_L i_c^{\,2} = 10^3(15 \times 10^{-4})^2 = 2.25\text{ mW}.$$

Maximum power output of transistor 1

In this case also, the maximum power which transistor 1 can supply is given by

$$P_{o1\max} = \tfrac{1}{4}\, E^2/\mathrm{h}_o.$$

but E cannot equated to the accepted value of $v_p = 10$ mV, because the load impedance Z_L is not large in relation to the output impedance h_o of the transistor.

This is because the output impedance is equal to:

$$Z_L = n^2\mathrm{h}_i = (\tfrac{1}{3})^2 \times 10^3 = 110\Omega.$$

We can calculate the value of E with the aid of the equivalent output circuit of transistor 1. The current through this circuit is:

$$i_c = V_{Z_L}/Z_L = 10^{-2}/110 = 9 \times 10^{-5}\ \mathrm{A} = 90\ \mu\mathrm{A}.$$

Now Z_L can be neglected in relation to h_o, so that the voltage E is given by:

$$E = i_c\mathrm{h}_o = 90 \times 10^{-6} \times 10^4 = 0.9\ \mathrm{V}.$$

It follows that

$$P_{o1\max} = \tfrac{1}{4}E^2/\mathrm{h}_o = \tfrac{1}{4}(0.9)^2/10^4 = 2 \times 10^{-5}\ \mathrm{W} = 20\ \mu\mathrm{W}.$$

Power gain

In this case (current drive) the power gain is therefore:

$$G_P = P_{o2}/P_{o1\max} = 2250/20 = 102.$$

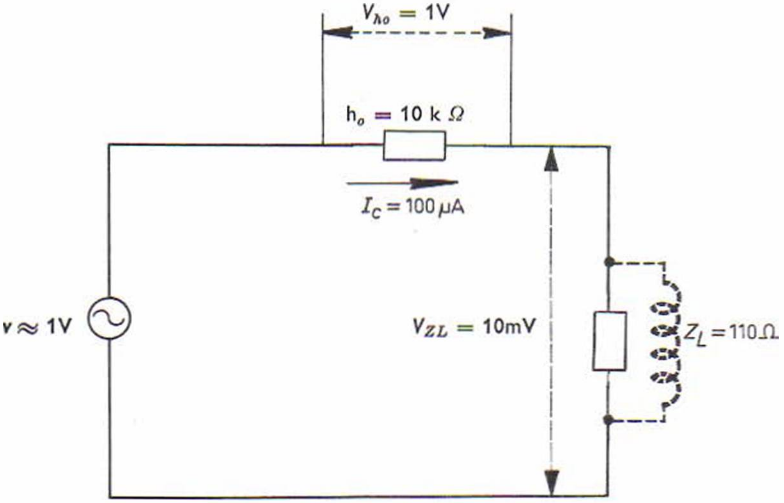

$V_{ho} = 1V$
$h_o = 10\ k\Omega$
$I_C = 100\ \mu A$
$v \approx 1V$
$V_{ZL} = 10mV$
$Z_L = 110\Omega$

Fig. 236

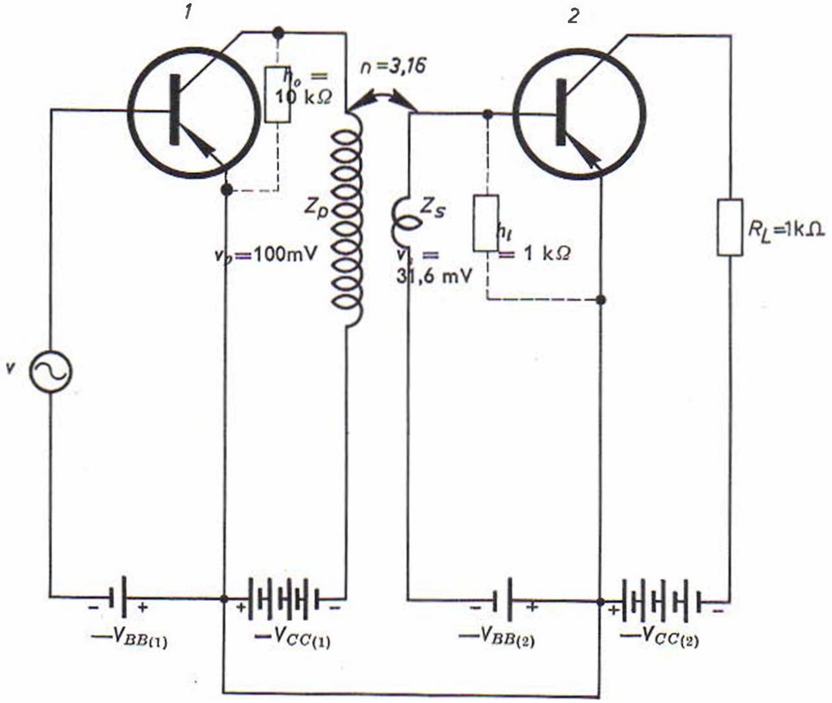

1
2
n = 3,16
h_o = 10 kΩ
Z_p
Z_s
v_p = 100mV
v_s = 31,6 mV
h_i = 1 kΩ
R_L = 1kΩ
v
−V_BB(1)
−V_CC(1)
−V_BB(2)
−V_CC(2)

Fig. 237

Matched drive

We will now consider the case when matched drive is applied (see Fig. 237). The two transistors are now coupled by a transformer of which the turns ratio n is chosen so that the output impedance of transistor 1, transformed to the secondary, is equal to the input impedance of transistor 2 (thus $h_o/n^2 = h_i$), from which it follows that:

$$n = \sqrt{h_o/h_i} = \sqrt{10^4/10^3} = 3.16$$

When transformed to the primary, the input impedance h_i of transistor 2 is of course also equal to the input impedance of transistor 1. We will now recalculate the power available at the output, and the maximum output power of transistor 1, in order to arrive at the value of the power gain.

Power available at the output

The power available at the output is:

$$P_{o2} = R_L\, i_c^2 = 1000\, i_c^2.$$

The value of i_c depends on the base current, which in turn is determined by the voltage across the secondary of the transformer, and by the input impedance of transistor 2. Let us assume that in this case the voltage v_p across the primary of the transformer is 0.1 V so that the voltage across the secondary is:

$$v_s = v_p/n = 0.1/3.16 \backsimeq 31.6 \text{ mV}.$$

This voltage means that the current flowing in the input circuit of transistor 2 is:

$$i_b = v_s/h_i = 31.6 \times 10^{-3}/10^3 = 31.6 \times 10^{-6} \text{ A} = 31.6\ \mu\text{A}.$$

If we again assume that the gain of transistor 2 is equal to 50, it follows that the collector current is:

$$i_c = 50 \times 31.6 = 1580\ \mu\text{A}.$$

The power available at the output is thus given by:

$$P_o^2 = R_L i_c^2 = 10^3 \times (1580 \times 10^{-6})^2 = 25 \times 10^{-4} \text{ W} = 2500\ \mu\text{W}.$$

Maximum output power of transistor 1

The maximum power which transistor 1 can supply can be calculated in this case from the expression:

$$P_{o1\mathrm{max}} = \tfrac{1}{4}E^2/\mathrm{h}_o$$

since the load impedance, that is, the input impedance of transistor 2, when transformed to the primary, is:

$$Z_L = 1000 \times 3.16^2 = 10\,000\ \Omega$$

which equals the output impedance h_o of transistor 1 (see Fig. 238). Since Z_L and h_o are both 10 kΩ, E is now given by:

$$E = 2v_p = 2 \times 0.1 = 0.2\ \mathrm{V}.$$

from which it follows that the maximum power which transistor 1 can supply is:

$$P_{o1\mathrm{max}} = \tfrac{1}{4} \times 0.2^2/10.000 = 10^{-6}\ \mathrm{W} = 1\ \mu\mathrm{W}.$$

Power gain

The power gain in this case (matched drive) is therefore:

$$G_P = P_{o2}/P_{o1\mathrm{max}} = 2500/1 = 2500.$$

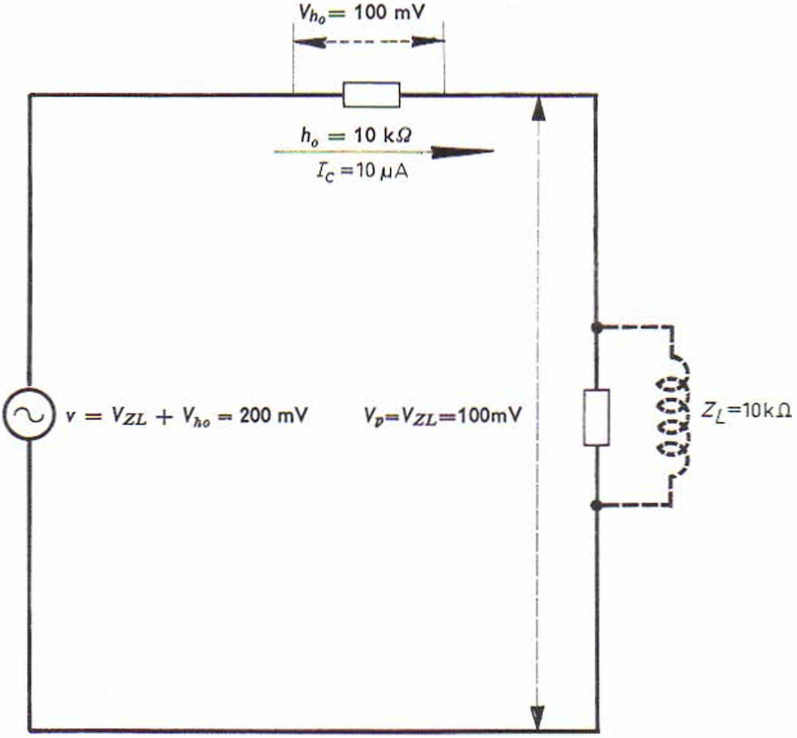

$V_{h_o} = 100$ mV
$h_o = 10$ kΩ
$I_C = 10$ μA
$v = V_{ZL} + V_{ho} = 200$ mV
$V_p = V_{ZL} = 100$mV
$Z_L = 10$kΩ

Fig. 238

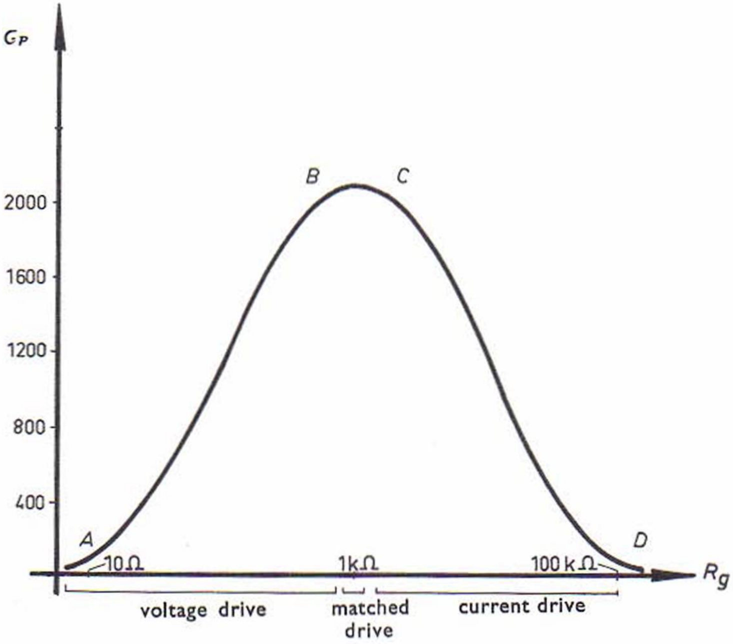

G_P
2000
1600
1200
800
400
B
C
A
D
10 Ω
1 kΩ
100 kΩ
R_g
voltage drive
matched drive
current drive

Fig. 239

25.3. Summary

Power gain

The above calculations show that the power gain depends very much on the method of driving the transistor. In Fig. 239 the gain is plotted as a function of the generator impedance. Region *AB* relates to low generator impedances in which the transistor is therefore voltage driven. At very low values of the generator impedance the power gain is almost zero, but increases quickly as the impedance increases. The region *BC* relates to the conditions obtaining when the generator impedance is equal or practically equal to the impedance of the transistor, so that matched drive is being applied. The power gain reaches its maximum value in this region. The region *CD* relates to generator impedances which are greater than the input impedance of the transistor, so that the transistor is current driven. In this region the power gain decreases rapidly with increasing value of the generator impedance.

The above considerations are not affected by the shape of the $-I_C = f(-I_B)$ characteristic.

Distortion

When we refer to distortion as a function of the generator impedance we must distinguish between the following two cases:

Transistor driven by small signals.
Transistor driven by large signals.

As already explained, the non-linear distortion caused by the curvature of the characteristic can be neglected if the input signal is of small amplitude. This is because only a very small portion of the characteristic is used in this case, and this small portion can be regarded as practically straight.

In this case the distortion caused by the transistor amplifier is practically constant, and is only slight, whatever the method of driving.

On the other hand however, if the transistor is driven by large signals, the distortion will in fact be determined by the form of the $-I_C = f(-I_B)$ characteristic.
If the operating conditions of the transistor have been selected to supply a small output power, and the straight portion of the characteristic is being utilised, the curve showing the distortion plotted against the generator impedance will be as in Fig. 240. The distortion will be high with voltage current drive, will decrease with matched drive, and will tend to a minimum value for current drive.

If the transistor is being driven by large signals and its working point has been chosen to supply large output powers, so that the curved portion of the $-I_C = f(-I_B)$ characteristic is being utilised, the curve of distortion plotted against the generator impedance will be as shown in Fig. 241. In this case the distortion increases with the value of the generator impedance, and has its minimum value for voltage drive.

Selection of the best compromise

We will now investigate which method of driving offers the best compromise in relation to power gain and distortion. As we have already seen, the operation of an amplifier stage is controlled by the following important factors:

The power gain of the stage.
The distortion arising in the stage.

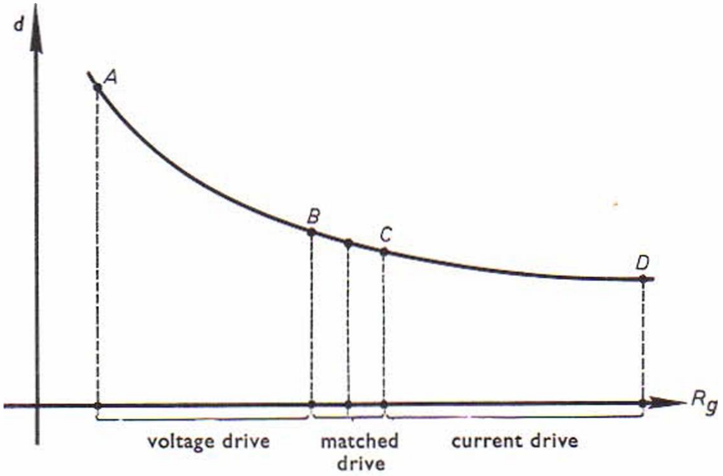


Fig. 240

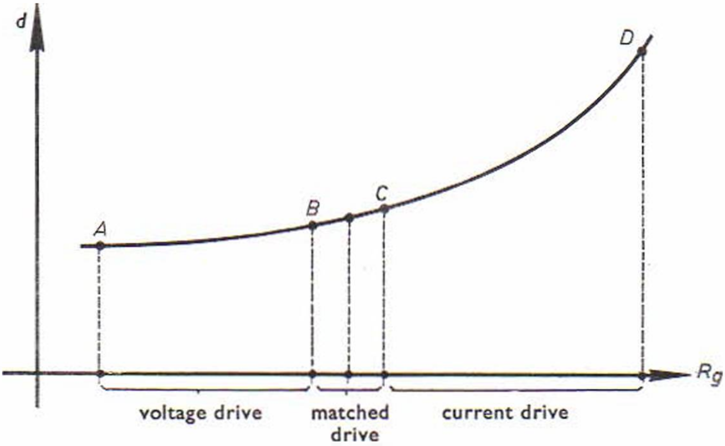


Fig. 241

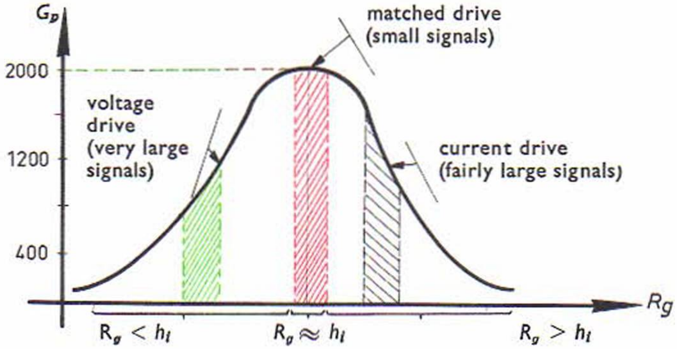


Fig. 242

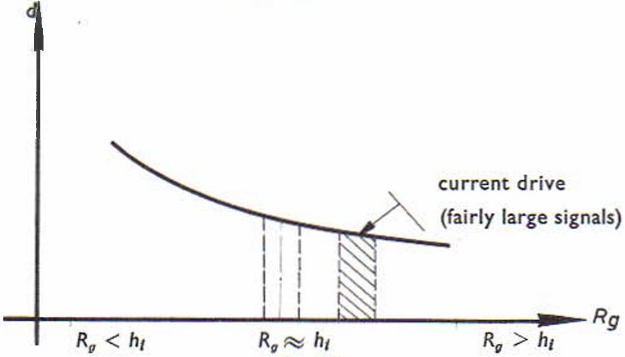


Fig. 243

($-I_C = f(-I_B)$ = linear)

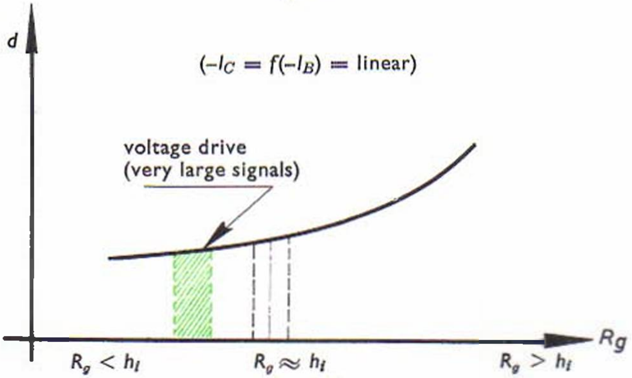


Fig. 244

($-I_C = f(-I_B)$ = curved)

An amplifier is required to give the greatest possible amplification with the smallest possible distortion. We must make a distinction here between the amplification of small and large signals, while in the latter case we can make a further distinction between fairly large and very large signals.
For amplifying small signals, the largest amplification is obtained with matched transistor drive (Fig. 242). The distortion arising in this case is practically constant and extremely small, independent of the method of driving. This means that it is an advantage, for the amplification of small signals, to employ only matched drive (red shaded area).

For amplifying fairly large signals, we must look for a compromise between maximum power gain and minimum distortion. As in the previous case, the power gain is a maximum with matched drive (Fig. 242), but the distortion will be too large, although it decreases if current drive is utilised (see Fig. 243). This means that the use of current drive is recommended for the amplification of fairly large signals but, in order to prevent a drastic reduction in power gain, the value of the generator impedance must not be excessively high (black shaded area).

For the amplification of very large signals too, the largest power gain is of course obtained with matched drive. In this case the distortion is the greatest for current drive, but is still appreciable with matched drive (see Fig. 244). For this reason, it is recommended that voltage drive should be employed but, in order to keep a large power gain, the generator impedance must not be too low (green shaded area).

CHAPTER 26

Transistor bias voltages

The object of applying a bias voltage is to fix the working point of the transistor in the absence of an input signal. Let us draw a load line in the $-I_C = f(-V_{CE})$ characteristic of Fig. 245, so that this line cuts the abscissa at 10 V (point A); the slope of the load line is given by $\tan A = 1/R_L$. We must now determine the point on this load line which will define the collector current in the absence of an input signal.

In order to obtain the greatest possible output signal with the lowest possible distortion, the swing of the signal along the load line must be as great as possible and must have the same amplitude on both sides of the working point. The working point must consequently be chosen such that the emitter voltage V_{CE} (in the absence of the input signal) is equal to half the supply voltage $-V_{CC}$, in this case therefore, such that (see Fig. 246):

$$V_{CE} = V_{CC}/2 = 10/2 = 5 \text{ V}.$$

Fig. 247 shows the equivalent circuit of the transistor. The voltage across the load resistance R_L is equal to the difference between the supply voltage and the collector-emitter voltage:

$$V_{RL} = V_{CC} - V_{CE} = 10 - 5 = 5 \text{ V}.$$

Since the current $-I_C$ causes a voltage drop of 5 V across the load resistance of 1 kΩ we can calculate the value of this current:

$$I_C = V_{RL}/R_L = 5/1000 = 5 \times 10^{-3} \text{ A} = 5 \text{ mA}.$$

The projection of the point $-V_{CE} = 5$ V (point B) determines point C on the characteristic and the projection of this point on the $-I_C$ axis determines point D, which of course corresponds to a collector current $-I_C$ of 5 mA.

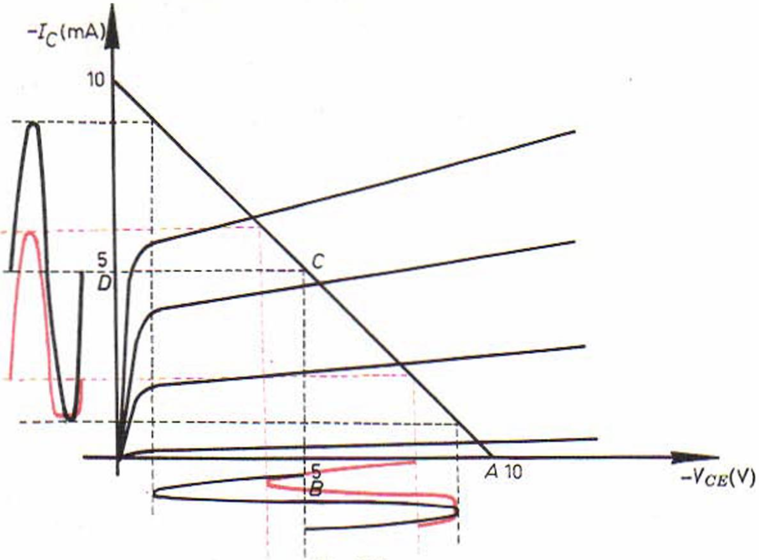


Fig. 245

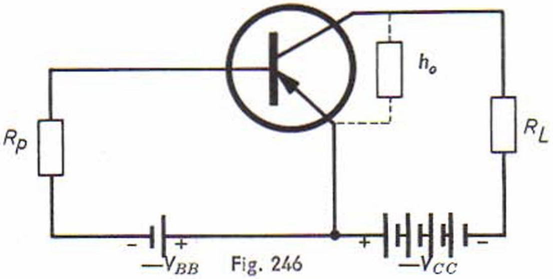

h_o
R_p
R_L
-
+
$-V_{BB}$
+
-
$-V_{CC}$

Fig. 246

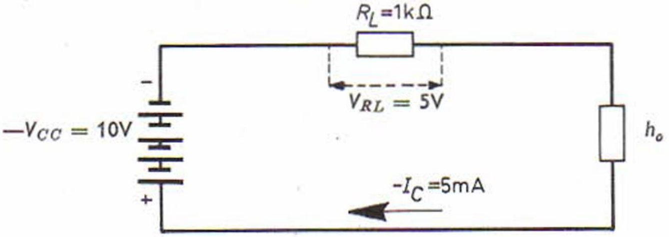

R_L=1kΩ
V_{RL} = 5V
-
$-V_{CC}$ = 10V
+
h_o
$-I_C$=5mA

Fig. 247

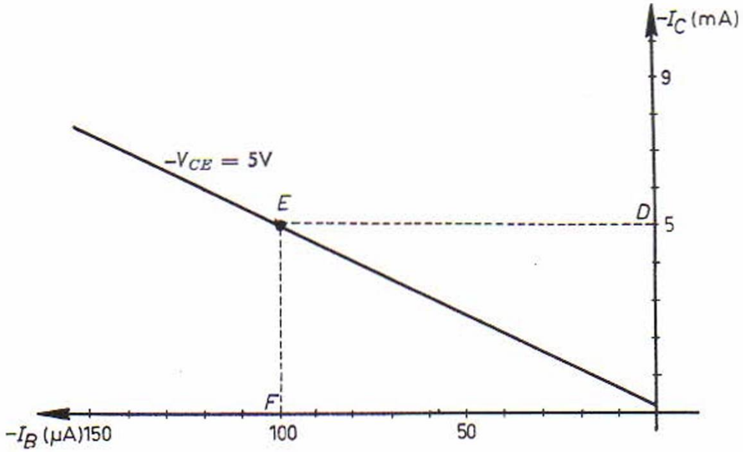


Fig. 248

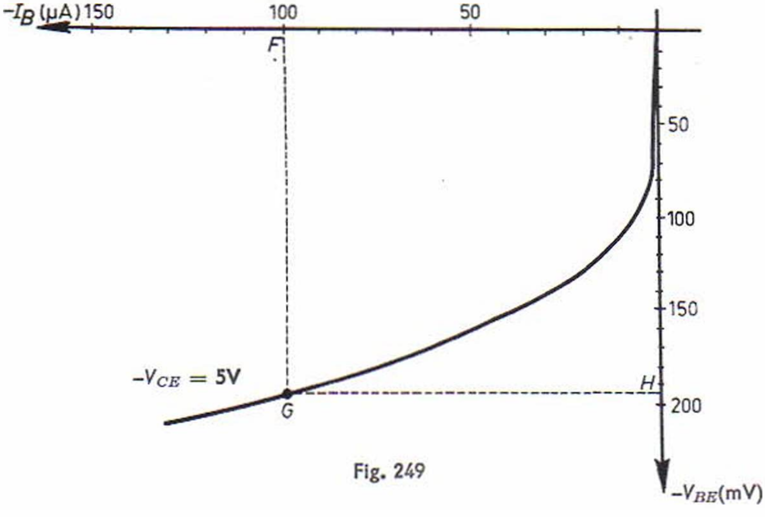


Fig. 249

Suppose that the current gain of the transistor h_{FE} is equal to 50. Neglecting the leakage current $-I_{CEO}$, we can equate the collector current to:

$$-I_C = h_{FE} \cdot (-I_B)$$

from which it follows that the constant base current of the transistor is:

$$I_B = I_C/h_{FE} = 5/50 = 0.1 \text{ mA}.$$

Once the constant collector current $-I_C$ is known, the constant base current can be read directly from the $-I_C = f(-I_B)$ characteristic. (see Fig. 248). The projection of point D ($-I_C = 5$ mA) defines point E on the characteristic and the projection of this point on the $-I_B$ axis determines point F, corresponding to 100 μA.

Let us now examine the $-I_B = f(-V_{BE})$ characteristic of the transistor. (see Fig. 249). The projection of point F on this characteristic gives point G, corresponding to point H on the $-V_{BE}$ axis, from which it follows that the base-emitter voltage (in the absence of an input signal) is equal to $-V_{BE} = 195$ mV. It is therefore possible to deduce the bias voltage of the transistor from the base-emitter voltage $-V_{BE}$, but as the examples show, it is a simpler matter to determine the constant base current $-I_B$ if the current gain of the transistor is known. The graphical method is certainly the quickest, while the calculation must assume that the $-I_C = f(-I_B)$ characteristic is absolutely straight. This is because the current gain h_{FE} is of course the factor giving the relationship between the collector current and the base current, and it must be accurately known, which is not always the case.

26.1. Graphical determination of the working point of a transistor

Fig. 250 shows all the normal transistor characteristics. On the $-I_C = f(-V_{CE})$ characteristic we draw the load line whose slope corresponds to a low resistance R_L of 1 kΩ. We select a working point on this characteristic such that $-V_{CE} = 5$ V, at which value the transistor can in fact supply the maximum power at the minimum distortion. At point B we erect the perpendicular to cut the load line at C. The projection of this point determines point D on the $-I_C$ axis, corresponding to a constant collector current $-I_C$ of 5 mA. The continuation of line CD intersects the $-I_C = f(-I_B)$ characteristic at point E and the projection of this point on the $-I_B$ axis determines point F, corresponding to a constant base current $-I_B$ of 100 µA. The continuation of line EF intersects the $-I_B = f(-V_{BE})$ characteristic at point G, whose projection on the $-V_{BE}$ axis is represented by point H. This last point gives the base-emitter voltage of the transistor in the absence of an input signal. In this way we have used the group of characteristics to determine the working point on the $-I_B = f(-V_{BE})$ characteristic, and we have now merely to draw the input load line in this graph, in order to find out how it moves as a function of variations in the input current.

Consequently, the bias voltage which is applied proves to be of the greatest importance for the proper operation of a transistor as amplifier. We will now investigate how the constant base current can be adjusted in practice. We shall see that the bias can be obtained in two ways:

By means of a bias battery ($-V_{BB}$).
By means of the collector supply voltage $-V_{CC}$.

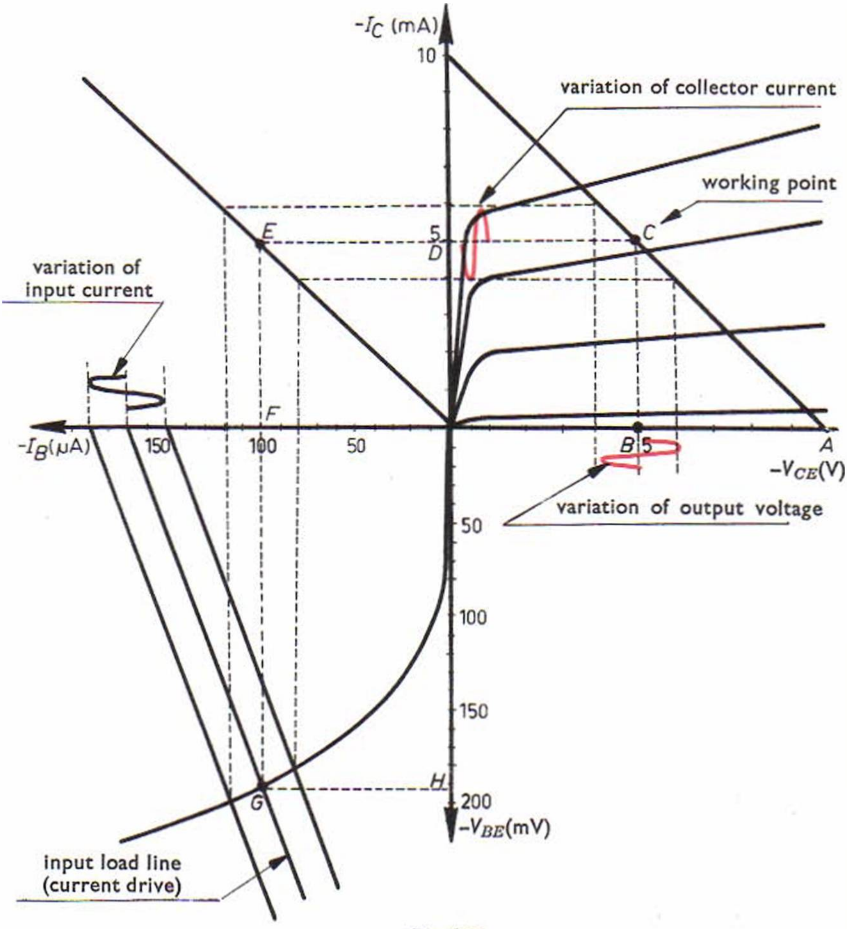

-I_C (mA)
10
variation of collector current
working point
E
5
D
C
variation of input current
F
-I_B(μA)
150
100
50
B
5
A
-V_{CE}(V)
variation of output voltage
50
100
150
H
200
G
-V_{BE}(mV)
input load line
(current drive)

Fig. 250

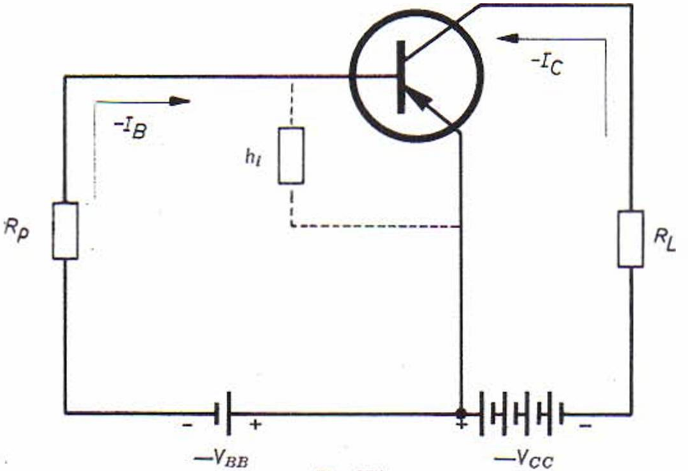

$-I_C$
$-I_B$
h_I
R_p
R_L
-
+
+
-
$-V_{BB}$
$-V_{CC}$

Fig. 251

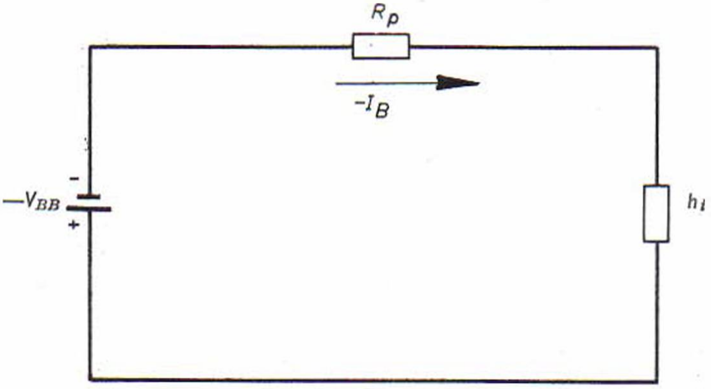


Fig. 252

26.2. The use of a bias battery

When a separate bias battery is used, we obtain the circuit of Fig. 251. Assuming that the collector supply voltage $-V_{CC}$ is 10 V, the collector-emitter voltage $-V_{CE}$ is 5 V, corresponding to a collector current $-I_C$ of 5 mA. The graphical derivation on the previous page showed that a constant collector current of 5 mA meant that the transistor in question had a constant base current $-I_B$ of 100 μA.

Fig. 252 is the equivalent circuit diagram of the input circuit of the transistor. This consists of a battery ($-V_{BB} = 1.5$ V) connected in series with the bias resistor R_p and the input resistance of the transistor. If we denote the sum of these resistances by R, we can use Ohm's law to determine the value of this resistance from the current which has to flow through the circuit (the constant base current):

$$R = V_{BB}/I_B = 1.5/10^{-4} = 15\ \text{k}\Omega.$$

Since the input impedance of the transistor is of the order of 1 kΩ this resistance can be neglected compared with the total resistance of the circuit, so that R_p can be taken as 15 000 Ω. This shows that the bias resistance (the base resistor) which must be connected in series with the battery $-V_{BB}$ must have a value of 15 kΩ.

26.3. Bias voltage taken from the supply battery

The circuit which has just been described requires an extra battery to supply the bias voltage: it is simpler if the battery which supplies the collector is also used for this purpose. Suppose that the supply voltage is $-V_{CC} =$ 10 V, that the load resistance in the collector circuit has a value of 1 kΩ and that the collector-emitter voltage in the absence of a signal is 5 V, so that the constant collector current $-I_C$ is equal to 5 mA.

As we found graphically for this type of transistor, a collector current of 5 mA corresponds to a base current $-I_b$ of 100 μA. Fig. 254 shows the equivalent circuit diagram of the input circuit. This circuit consists of the supply battery having a voltage $-V_{CC} = 10$ V, with two impedances in series; the bias resistor R_p and the input impedance h_i of the transistor. The total resistance of the circuit is equal to the sum of these two resistances.

$$R = R_p + h_i.$$

The value of the resistance R can be deduced from the voltage across the terminals of the circuit and the current which must flow through it, $(-I_B = 100\ \mu\text{A})$:

$$R = V_{CC}/I_B = 10/10^{-4} = 10^5 = 100\ \text{k}\Omega.$$

The input impedance (of the order of 1 kΩ) can be neglected in relation to R, so that the bias resistor can be taken as having a resistance of 100 kΩ. The required bias is thus obtained by connecting a 100 kΩ resistor between the negative pole of the supply battery and the base of the transistor. The value of this resistance is therefore determined by the constant base current and by the value of the supply voltage.

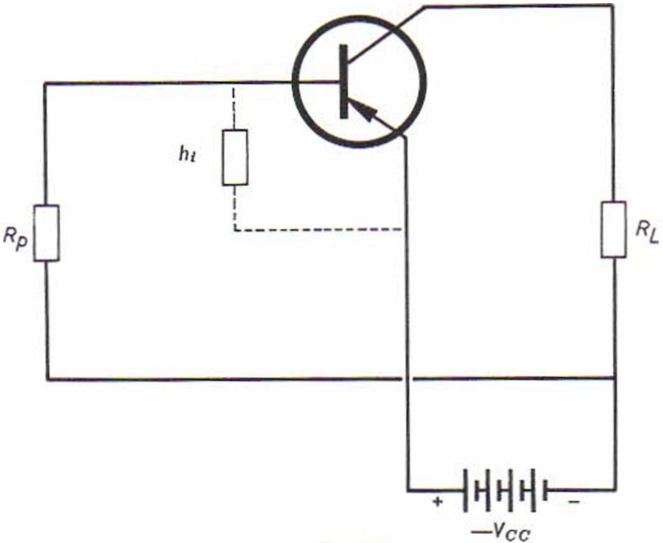


Fig. 253

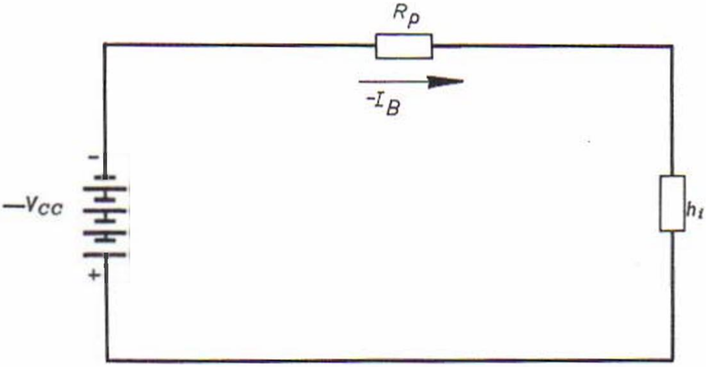


Fig. 254

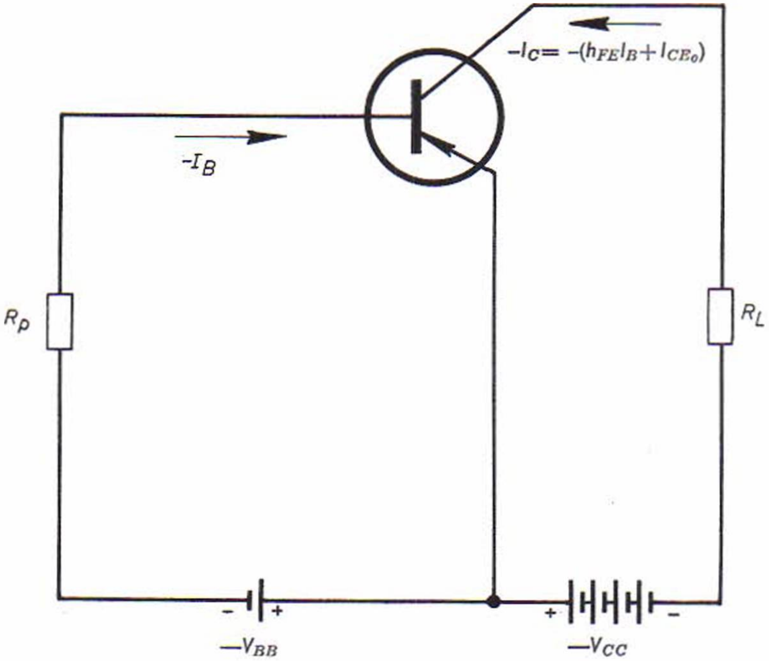

$-I_C = -(h_{FE}I_B + I_{CEo})$
$-I_B$
R_p
R_L
$-$
$+$
$-V_{BB}$
$+$
$-$
$-V_{CC}$

Fig. 255

CHAPTER 27

Thermal stability

The temperature of the collector-base junction and of the base-emitter junction must be prevented from exceeding a given value, because this could damage the transistor. With reference to the collector-base junction, several precautions must be taken. Suppose that the temperature of this junction is Tj. This depends on three factors, namely:

The ambient temperature T_{amb}
The collector dissipation P_C
The thermal resistance K between the junction and its surroundings.

The relationship between Tj and these three factors is given by the expression:

$$Tj = T_{\text{amb}} + KP_C.$$

27.1. Ambient temperature

By ambient temperature we understand the temperature of the medium in which the transistor is situated. The temperature of the collector junction increases with the ambient temperature, so that this factor is of the greatest importance in considering the operation of transistors.

27.2. Collector dissipation

The power dissipated in the collector is given by the product of the collector-emitter voltage and the collector current:

$$P_C = V_{CE}I_C.$$

Now the collector current is equal to the product of the base current and the current amplification factor, plus the leakage current $-I_{CEO}$ (see Fig. 255):

$$I_C = h_{FE}I_B + I_{CEO}.$$

The leakage current increases sharply with increasing temperature, so that an increase in the ambient temperature is accompanied by an increase in the collector current.

27.3. Thermal resistance

We can compare the effect of thermal resistance with that of electrical resistance. To do this, we will commence with the circuit of Fig. 256, in which the capacitor *C* can be connected either to the terminals of a battery (of output voltage *V*) or across the resistor *R*. If switch 1 is closed and switch 2 is open (printed black), the capacitor quickly becomes charged by the battery, since the battery resistance is negligible. If switch 1 is now opened and switch 2 closed (printed red) the capacitor will discharge via resistor *R*. The lower the resistance of this resistor, the more rapidly will the energy stored in the capacitor flow away. If *R* equals 0, the capacitor will discharge immediately.

We can now regard the actual transistor material as a thermal "capacitor". The temperature reached by this material can be regarded as the voltage to which the capacitor of Fig. 256 becomes charged, while its thermal capacitance can be compared to the capacitance of the capacitor. The energy stored in the transistor material flows out towards its surroundings in the form of a thermal current, at a rate which depends on the thermal resistance it encounters, just as the energy stored in the capacitor flows away in the form of electric current via the resistor *R*, at a rate which is determined by the resistance of this resistor.

Transistors for larger powers are fitted with cooling fins which reduce the thermal resistance, so that the power developed in these transistors can easily be transferred to the surroundings. This state of affairs is shown in Fig. 257, in which the maximum permissible dissipation P_C of an OC72 transistor is plotted against the ambient temperature T_{amb}. It will be seen that at an ambient temperature of 25° C the collector may dissipate a maximum power of 125 mW if the transistor is not fitted with a cooling fin (printed black). When a cooling fin is fitted (red line), the transistor can dissipate a power of 165 mW at the same ambient temperature.

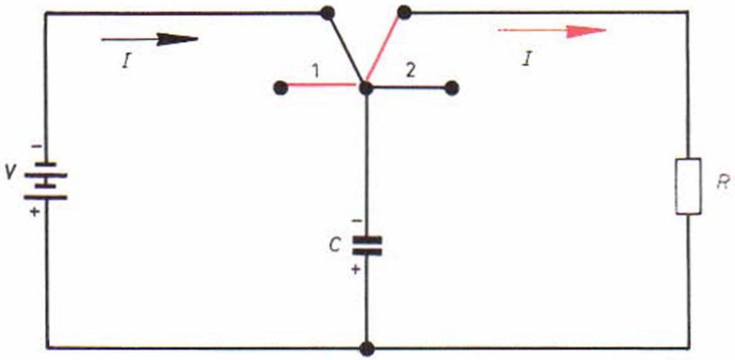


Fig. 256

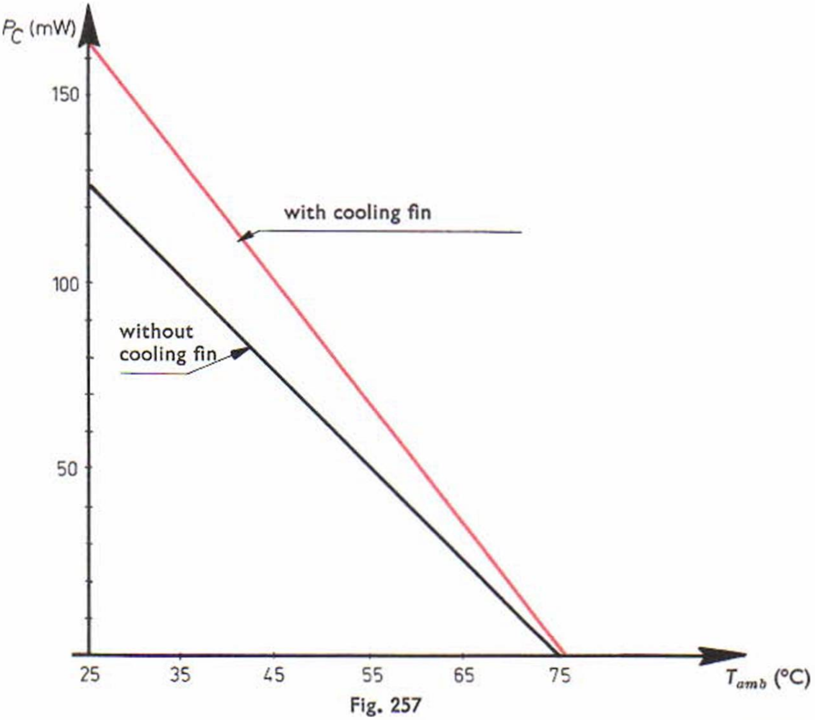


Fig. 257

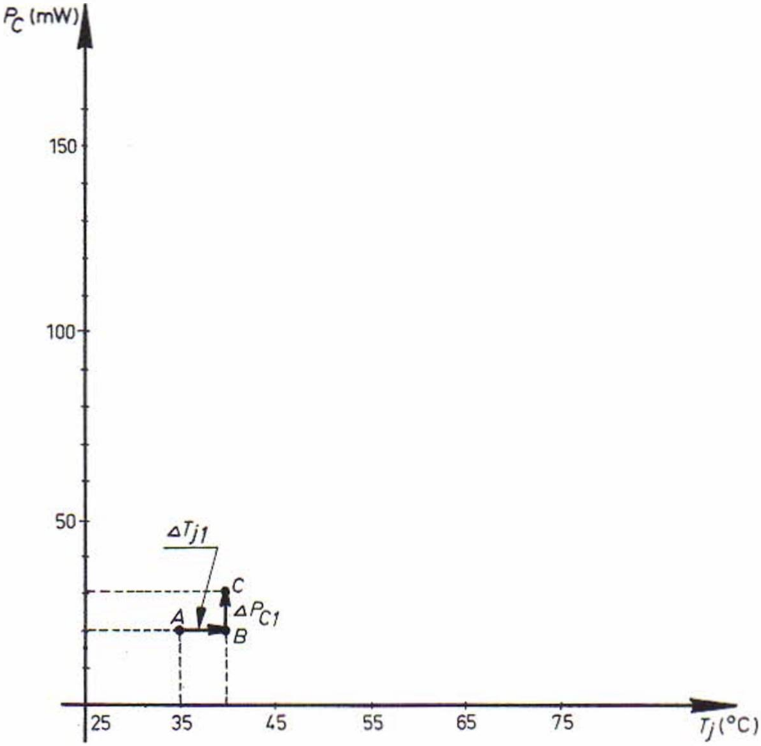


Fig. 258

We will now examine the junction temperature as a function of the power dissipated in the collector. To this end we plot the temperature Tj as a function of the collector dissipation P_C (Fig. 258). Suppose that the transistor is working at a given instant with a collector dissipation of 20 mW and that $Tj = 35°$ C (point A).

If the ambient temperature increases, this will result in an increase of Tj $(= T_{\text{amb}} + KP_C)$. Point A on the characteristic moves parallel to the abscissa, the extent of this movement being determined by the increase in Tj. Suppose that $Tj_1 = 40°$ C. The difference between the junction temperature at point B and that at point A is termed the temperature increase ΔTj_1, so that in this case:

$$\Delta Tj_1 = 40 - 35 = 5 \text{ deg. C.}$$

This increase in temperature results in an increase in the leakage current $-I_{CEO}$. (As explained earlier, this current is caused by the breakage of bonds between germanium atoms, and we have seen that the number of broken bonds increases with the temperature). As a result of the increase in the leakage current $-I_{CEO}$ the collector current $I_C = \text{h}_{FE}I_B + I_{CEO}$, and with it the collector dissipation, will also increase. This increase ΔP_{C1} will show itself in a movement of the working point from B to C, and the increase in the dissipated power results in a further increase of Tj.

Let us indicate this temperature increase by ΔTj_2. This latest increase in temperature means that the working point moves from C to D (see Fig. 259), and this in its turn means that the leakage current $-I_{CEO}$ increases still further, so that the power dissipated in the collector increases once more. This further increase ΔP_{C_2} means that the working point moves again from D to E, and that there will be a further increase ΔTj_3 in the temperature of the base-collector junction (point F). In its turn this will lead to yet another increase in the power dissipated in the collector ΔP_{C_3} (point G). This cumulative effect proceeds until equilibrium is established.

We represent the relationship between the increase in the collector dissipation and the increase of the junction temperature which is the cause of this, by:

$$\lambda = \Delta P_C / \Delta Tj.$$

Now λ decreases in proportion as the temperature increases. Equilibrium is attained when the product $K\lambda$ becomes less than 1. (Where K represents the thermal resistance).

It is true that equilibrium will always be reached, but this may occur at a temperature which is higher than the maximum permissible value for the junction temperature. For germanium transistors equilibrium must always be reached at a temperature of not more than 75° C. The smaller the product $K\lambda$, the sooner will the equilibrium condition be reached, so that low values of K and of λ are desirable.

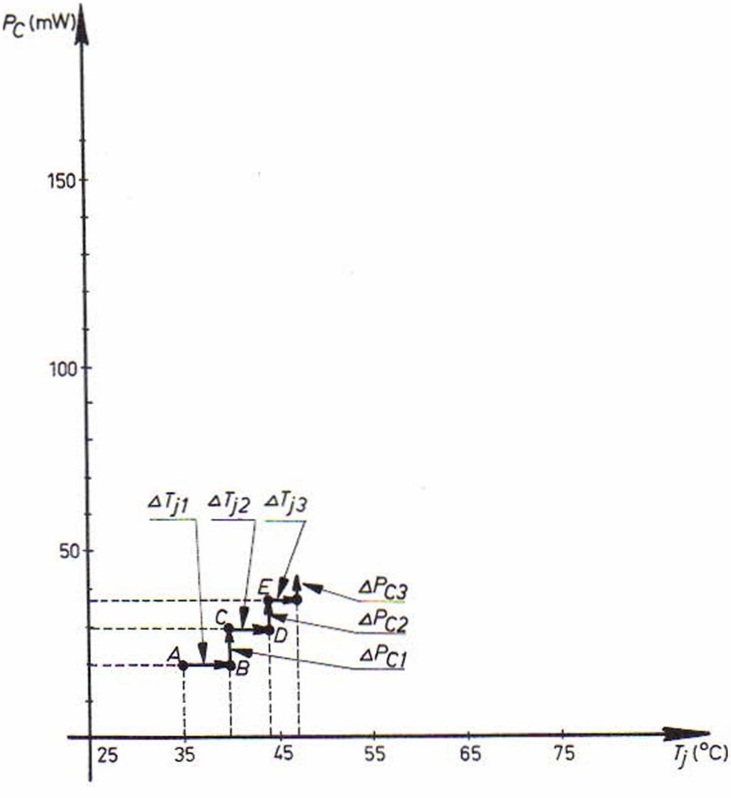

P_C (mW)
150
100
50
ΔT_{j1}
ΔT_{j2}
ΔT_{j3}
ΔP_{C3}
ΔP_{C2}
ΔP_{C1}
A
B
C
D
E
25
35
45
55
65
75
T_j (°C)

Fig. 259

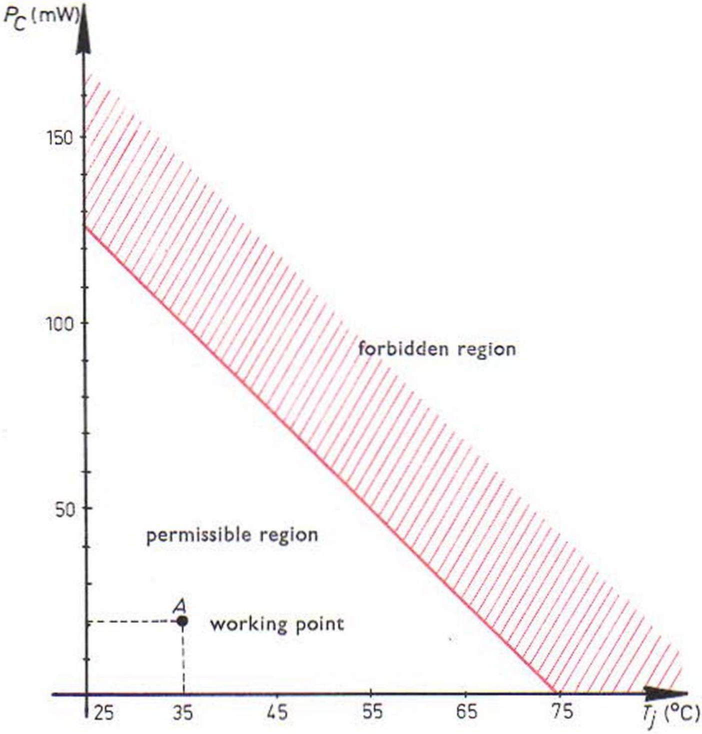

P_C (mW)
150
100
50
forbidden region
permissible region
A
working point
25
35
45
55
65
75
T_j (°C)

Fig. 260

27.4. Low thermal resistance

A reduction in the thermal resistance between the base-collector junction and the ambient surroundings can be recommended for two reasons:

Reduction of the product $K\lambda$.

Increase of the maximum power which may be dissipated in the collector.

Reduction of the product $K\lambda$

In order to prevent damaging a transistor, its working point as represented in Fig. 260 must never enter the region in which Tj is higher than the maximum permissible temperature (75° C, for example, for a germanium transistor), or the region in which the collector dissipation exceeds the prescribed maximum value. This means that equilibrium must be attained in the unshaded area of the figure. It is obvious that this condition will be reached more quickly if a rise in temperature resulting from an increase in the collector dissipation is limited by very rapid removal of the heat energy via a low thermal resistance. (See the electrical analogy on page 270).

Increase of the maximum collector dissipation

We can deduce the value of $P_{C\max.}$ from the equation $Tj = T_{\text{amb}} + KP_C$. From

$$Tj_{\max} = T_{\text{amb}} + KP_{C\max}$$

it follows that:

$$P_{C\max} = (Tj_{\max} - T_{\text{amb}})/K.$$

This shows that a decrease in the thermal resistance K is accompanied by an increase in the maximum permissible collector dissipation, or in other words, a reduction in the thermal resistance means that the collector current may be increased, so that the transistor can deliver a greater output power.

27.5. Low value of λ

Looking at the $P_C = f(Tj)$ characteristic of Fig. 261a, let us start from the working point corresponding to a collector dissipation of 20 mW at Tj = 35° C. (Point *A*) An increase of 5 deg C in the ambient temperature means that Tj increases by the same amount, so that the working point shifts from *A* to *B*.

ΔTj_1 represents the resulting increase in Tj. The direct result of this temperature increase is an increase in the collector current, and therefore in the collector dissipation P_C. This means that the working point shifts from *B* to *C*, corresponding to an increase of $\Delta P_{C_1} = 10$ mW in the collector dissipation. The result of this is that Tj increases further, by $\Delta Tj_2 = 4$ deg C. This increase in Tj corresponds to an increase in P_C of $\Delta P_{C_2} = 8$ mW.

Fig. 261b also shows an increase of 5 deg C in the ambient temperature corresponding to an increase $\Delta Tj_1 = 5$ deg C, whereby the working point moves from *A* to *B*. As λ is smaller than in the previous case, this temperature rise will cause a smaller increase in the collector current, and thus in the collector dissipation P_C.

In the case under consideration ΔP_{C_1} is only 5 mW (instead of 10 mW). The result is that the working point moves from *B* to *C*, but since ΔP_{C_1} is only half of ΔP_{C_1} in the previous case, ΔTj_2 will also be much less, for example 3 deg C. The increase in collector current due to ΔTj_2 and the resulting increase in collector dissipation ΔP_{C_2} are also less at the lower value of λ. For example $\Delta P_{C_2} = 2$ mW.

In these examples the product $K\lambda < 1$, so that in both cases we have a convergent series, and a stable condition is reached either outside the permissible area (Fig. 261a) or within this area (Fig. 261b). However, if $K\lambda > 1$, the successive steps will always increase in size, giving a divergent series which leads to the certain destruction of the transistor.

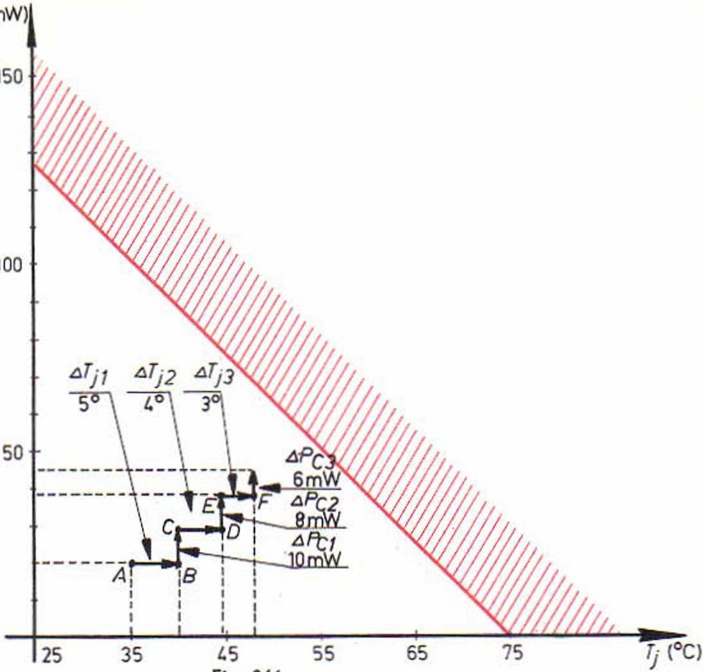


Fig. 261a

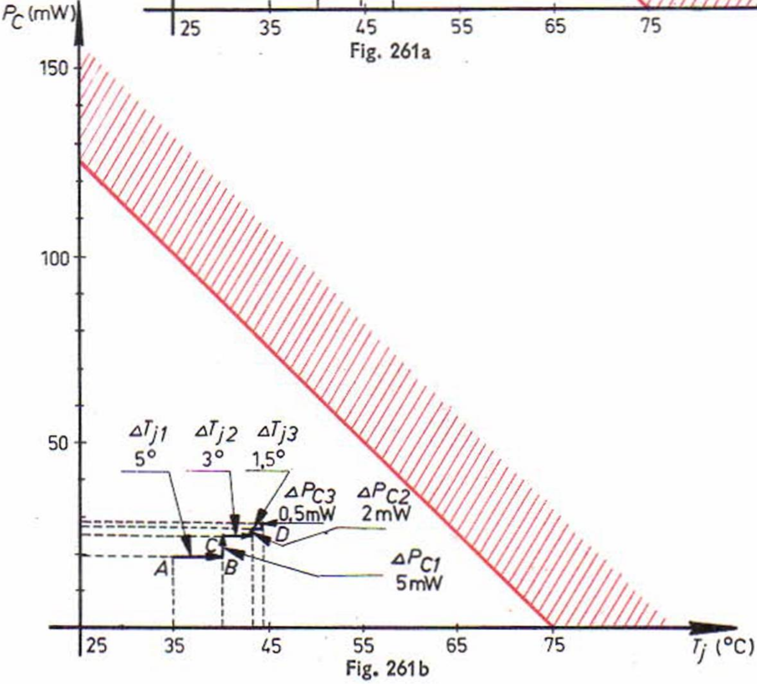


Fig. 261b

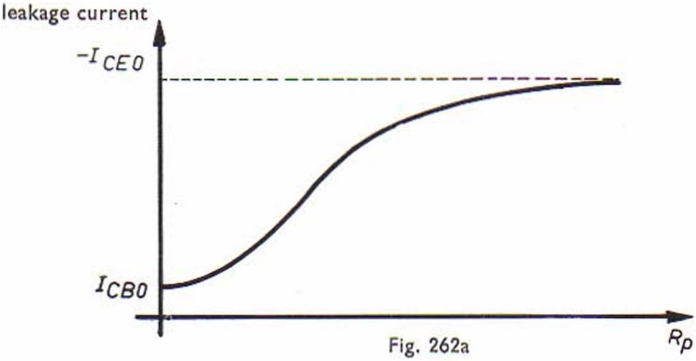


Fig. 262a

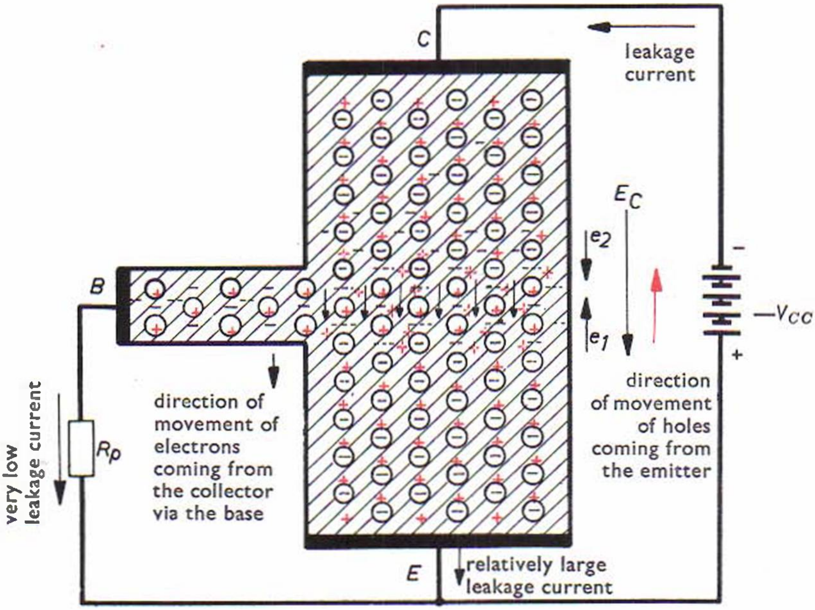

C
leakage current
B
E_C
e_2
e_1
$-V_{CC}$
-
+
very low leakage current
R_p
direction of movement of electrons coming from the collector via the base
direction of movement of holes coming from the emitter
E
relatively large leakage current

Fig. 262b

In the above, we have assumed that the increase in P_C is due solely to the increase in the leakage current $-I_{CEO}$, but in actual fact this leakage current is very dependent on the value of the generator impedance.

27.6. The relationship between the leakage current and the generator impedance

By definition, the leakage current $-I_{CEO}$ is not a useful characteristic of the transistor, since there is no practical use whatever in the possibility of using the transistor in common emitter with a floating base. In practice, the leakage current is understood to be the current flowing in the collector circuit of the transistor when the base is connected to the emitter via an impedance, i.e. the generator impedance R_g. This current is plotted as a function of R_g in Fig. 262a, in which figure we can distinguish two regions, depending on the value of R_g.

High generator impedance

With a high value of the generator impedance we obtain the case illustrated in Fig. 262b. The free electrons arising as a result of the breakage of bonds between germanium atoms in the collector diffuse to the base. These electrons can now follow one of two paths:

> They can cross the base-emitter junction under the influence of the external electric field E_C,
>
> They can flow through the generator impedance R_g.

If the generator impedance has a high value, most of the electrons will cross the base-emitter junction. As we showed in our consideration of the physical basis of the operation of transistors, a small flow of electrons between base and emitter will cause a much greater diffusion of holes from the emitter via the base to the collector.* It will thus be clear that the leakage current can become very considerable (much larger than the reverse current in the collector-base junction).

---

* Under these conditions the base-emitter junction becomes slightly biased in the forward direction.

Low generator impedance

Fig. 262c represents the situation when the generator impedance is very low (1 Ω). It is now easier for the electrons proceeding from the collector to flow through the impedance R_g than through the base-emitter junction. The exchange of electrons and holes between the base and the emitter is thus on a much smaller scale than in the previous case; as a result, the leakage current decreases considerably, and approaches the value of the reverse current of the collector-base junction. In practice, the leakage current for $R_g = 0$ will differ very little from $-I_{CBO}$.

We see from the curve of Fig. 262a that while the leakage current is considerable if the generator impedance or series resistance has a high value, this current and any variations of it may be neglected entirely with a low generator impedance. As a first approximation, one can say that by stabilisation of the base-emitter voltage of the transistor or the application of a constant direct voltage (small bias resistor) it should be possible to use the transistor without any danger of its temperature increasing. Under these conditions however, another effect occurs which must be ascribed to the thermal behaviour.

27.7. Effects connected with variations of the input characteristic resulting from increased temperatures

The base-emitter junction of a transistor which is connected in the forward direction can be represented as a diode, connected in the same direction. An increase in temperature causes the characteristic to shift in the direction of the $-I_B$ axis, which means that the base current increases slightly at a given base-emitter voltage. This is illustrated for a diode in Fig. 263a and for a transistor in Fig. 263b.

Since the leakage current and the effects it produces are reduced when the transistor is operating at constant voltage, we will assume that the bias in the circuit under consideration is supplied by a battery, so that the resistance of this circuit may be taken as zero (black input load line in Fig. 264). A rise in temperature will mean that the working point shifts from *A* to *B*, and from *B* to *C*. The projections of these three points (A', B' and C') determine the constant base current at the different temperatures; the variation of this current is not inconsiderable.

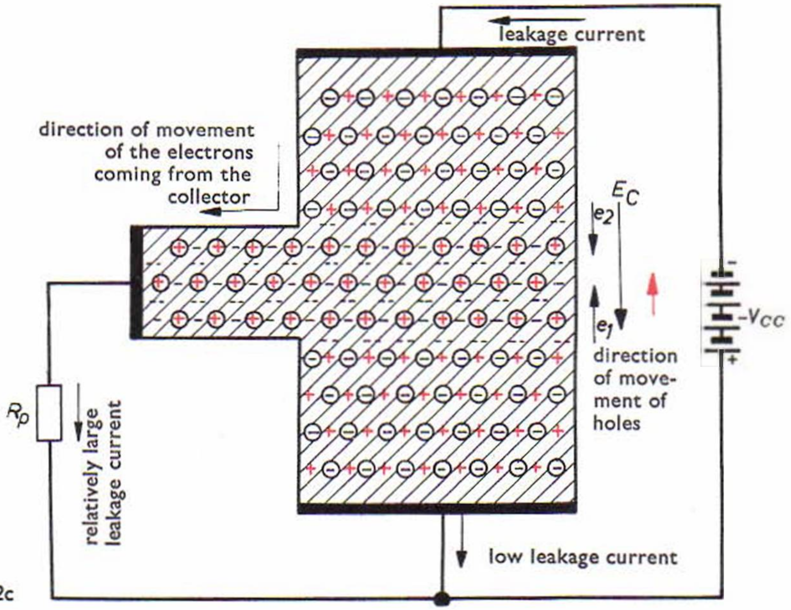
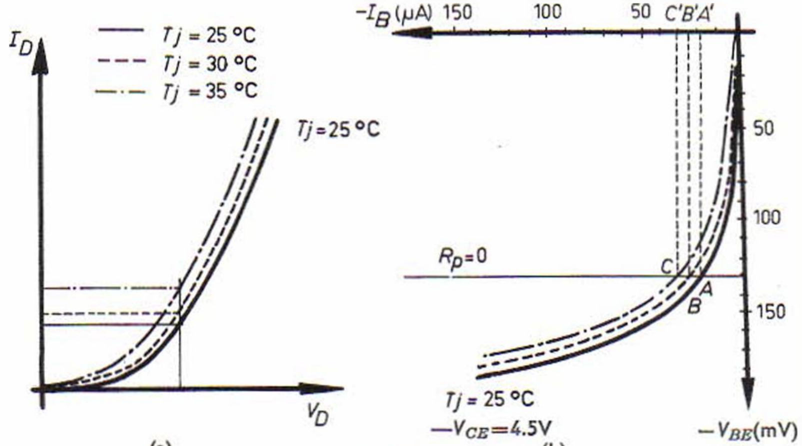

leakage current
direction of movement of the electrons coming from the collector
E_C
e_2
e_1
direction of movement of holes
$-V_{CC}$
-
+
R_p
relatively large leakage current
low leakage current

Fig. 262c



(a) (b)

Fig. 263

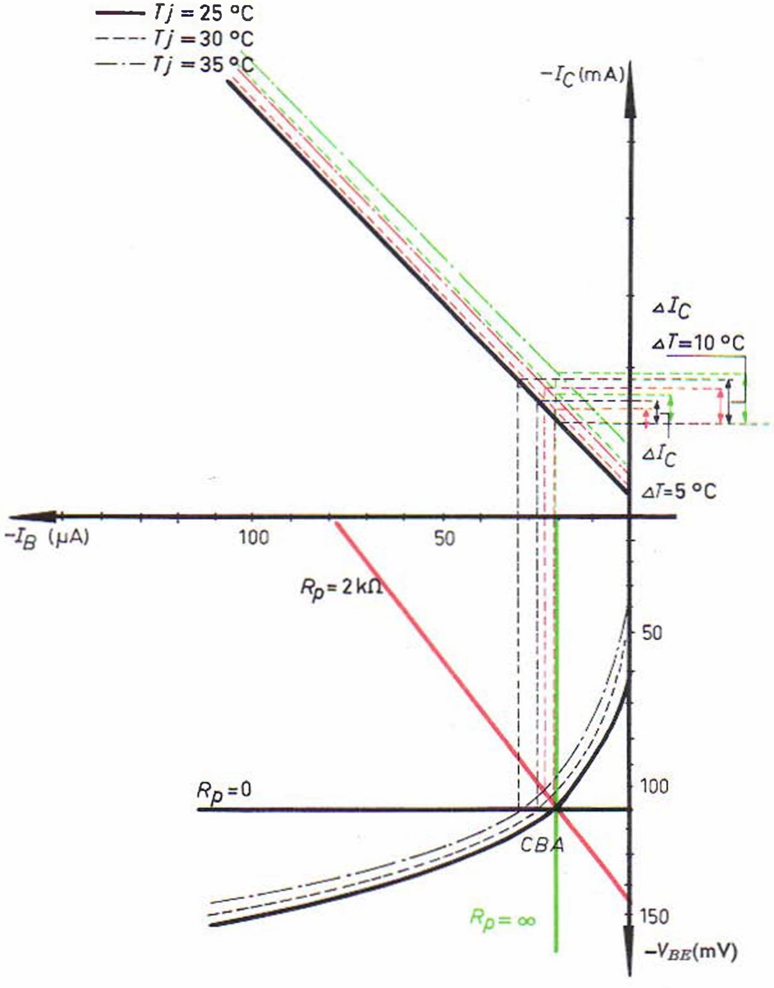

Tj = 25 °C
Tj = 30 °C
Tj = 35 °C
$-I_C$ (mA)
ΔI_C
ΔT=10 °C
ΔI_C
ΔT=5 °C
$-I_B$ (µA)
100
50
R_p= 2 kΩ
50
100
150
R_p = 0
C B A
$R_p = \infty$
$-V_{BE}$(mV)

Fig. 264

Since the collector current is equal to $h_{FE}I_B$ plus the leakage current, operation of the transistor with a constant voltage will, it is true, greatly reduce the effects of the leakage current, but will cause a not inconsiderable increase in the base current, and consequently a very substantial increase in the collector current.

If we connect a high resistance in series with the bias battery, the input load line will rotate on the working point (printed green in Fig. 264). We see that the changes in base current may then be neglected, but that those of the leakage current increase sharply.

Summarizing, therefore, we may say that a low series resistance will cause $h_{FE}I_B$ to fluctuate considerably with a negligible leakage current, while for a high series resistance the variations of $h_{FE}I_B$ will always be slight, but those of the leakage current will be appreciable. We must thus look for a compromise between these two extreme cases, since both have an unfavourable effect on the thermal behaviour of the transistor. In fact we must add to these effects that of the increase in the current gain of the transistor with increasing temperature. Under certain conditions this can become so large that it must be taken into account in designing a circuit.

Such a compromise is represented by the red input load line in Fig. 264. The transistor bias is obtained by means of a voltage divider in which the resistance values are so chosen as to give neither constant voltage operation (reduced effect of variations of the forward characteristic of the input diode), nor constant current operation (smaller leakage current). Although such stabilization is not perfect, it is often adequate. Because of its simplicity and cheapness, this circuit is often found in inexpensive circuits.

In the following pages we will discuss various circuits which give more effective stabilization of the transistor. For the sake of simplicity, the operation of these circuits will be explained with the aid of Ohm's law. Both the above forms of stabilization will be recognized in this discussion.

Every increase in $-I_C$ must be regarded as the result either of an increase of the leakage current at a high value of the generator impedance, or as an increase in $-I_B$ at a low value of this impedance. It is logical to stabilize the voltage first ($-V_{BE}$ = constant) in order to limit the leakage current, by using a voltage divider for feeding the base, and then to stabilize the current (compensation of variations in $-I_B$, for example, by connecting a resistor in series with the emitter). We will now consider the effect of temperature changes under various conditions.

Load formed by the primary of a transformer

In the circuit of Fig. 265a the transistor is loaded by a transformer whose primary impedance equals Z_L; the resistance R_Z can be neglected. Fig. 265b represents the equivalent output circuit of this transistor. The collector-emitter voltage $-V_{CE}$ is equal to the supply voltage, since the voltage drop across the primary of the transformer is negligible for direct current. An increase in the ambient temperature causes the collector current $-I_C$ to increase. Since the collector-emitter voltage $-V_{CE} = -V_{CC}$ remains constant, the collector dissipation P_C will also increase.

Consequently if the transistor is loaded by an inductance associated with a negligible resistance, there is no stabilization at all against variations of temperature.

Resistive load

Fig. 266a represents a transistor loaded by a resistance of value R_L. The equivalent output circuit is given in Fig. 266b. In this case, the collector-emitter voltage is:

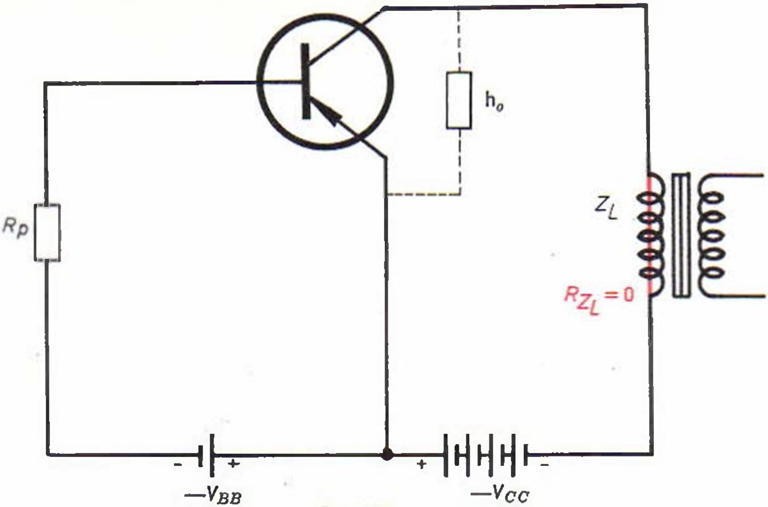
$$-V_{CE} = -V_{CC} - (-R_L I_C).$$

If the ambient temperature increases, the collector current I_C will increase, as a result of which the collector-emitter voltage decreases. This means that an increase in I_C is associated with a reduction in V_{CE}. The collector dissipation,

$$P_C = V_{CE} I_C$$

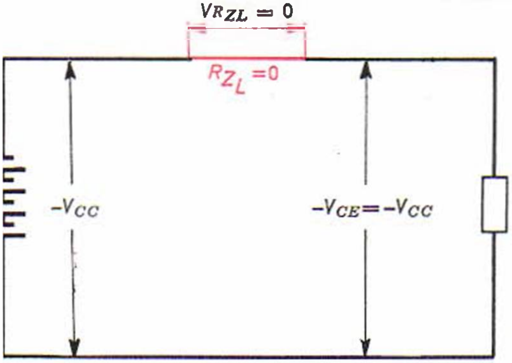
will thus change less than in the previous case for the same temperature increase.

Fig. 267 shows the collector dissipation plotted as a function of the collector current.



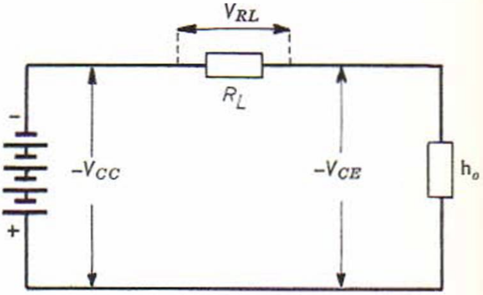
h_o
R_p
Z_L
$R_{Z_L} = 0$
- +
$-V_{BB}$
+ -
$-V_{CC}$

Fig. 265a



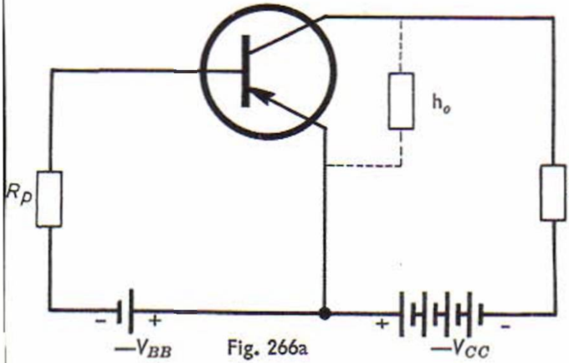
$V_{R_{ZL}} = 0$
$R_{Z_L} = 0$
$-V_{CC}$
$-V_{CE} = -V_{CC}$
h_o

Fig. 265b



V_{RL}
R_L
-
+
$-V_{CC}$
$-V_{CE}$
h_o

Fig. 266b



h_o
R_p
R_L
- +
$-V_{BB}$
+ -
$-V_{CC}$

Fig. 266a

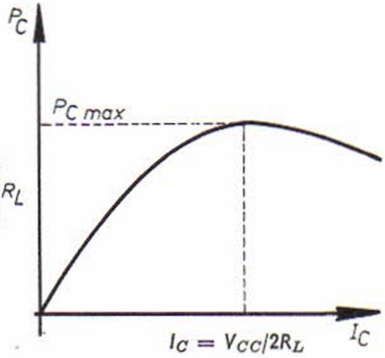


Fig. 267

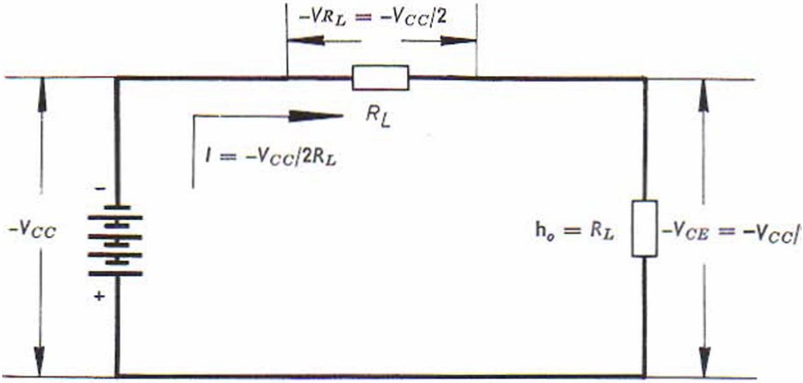

$-V_{R_L} = -V_{CC}/2$
R_L
$I = -V_{CC}/2R_L$
$-V_{CC}$
-
+
$h_o = R_L$
$-V_{CE} = -V_{CC}/$

Fig. 268

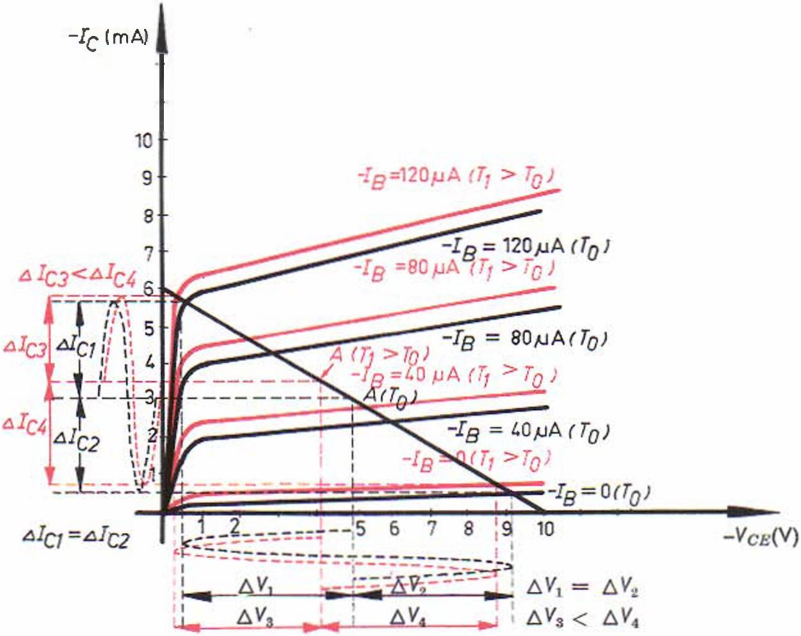


Fig. 269

Load consisting of a resistance equal to the output impedance of the transistor

Fig. 268 represents the special case in which the load impedance has the same value as the output impedance of the transistor. The collector current $-I_C$ is given by the quotient of the supply voltage $-V_{CC}$ and the total resistance of the circuit:

$$-I_C = -V_{CC}/(\mathrm{h}_o + R_L) = -V_{CC}/2R_L.$$

If the collector current is adjusted to this value, the circuit will always be stabilized against temperature variations. Of course, this method can only be applied if the load consists of a resistance. No account has been taken here of any spread between the collector currents of different transistors.

27.8. Stabilization of the working point of a transistor

The slope of the load line in the $-I_C = f(-V_{CE})$ characteristic of Fig. 269 is given by tan $A = 1/R_L$. The working point is chosen so as to give minimum distortion for maximum output power. With an increase in the temperature *Tj*, the collector dissipation and therefore the leakage current will increase, and there will be a danger of the working point shifting. Consequently, the larger the signal, the earlier will serious distortion occur.

Small signal amplifier

The load line shown in the $-I_C = f(-V_{CE})$ characteristic of Fig. 270 is for $R_L = 10\ k\Omega$. The working point is selected so that the collector-emitter voltage is equal to half the supply voltage $-V_{CC}$, so that $-V_{CE} = 5$ V in this case. The corresponding collector current $-I_C$ is given by the projection of the working point on the $-I_C$ axis, (point C), and is equal to 0.5 mA.

We will assume that the transistor is operating with small signals, that is, the variations of collector current which are caused by variations in the base current, have a very small amplitude. (For example, a peak-to-peak value of 40 µA for i_c). In this case a very large increase in the current $-I_C$ will be necessary to cause a change in the variations of the collector current or a movement of the working point.

It is thus extremely unlikely that the proper operation of a small signal transistor will be disturbed (red characteristic in Fig. 270). In most cases, it will even be possible to use the transistor with a constant collector current lower than the leakage current $-I_{CEO}$.

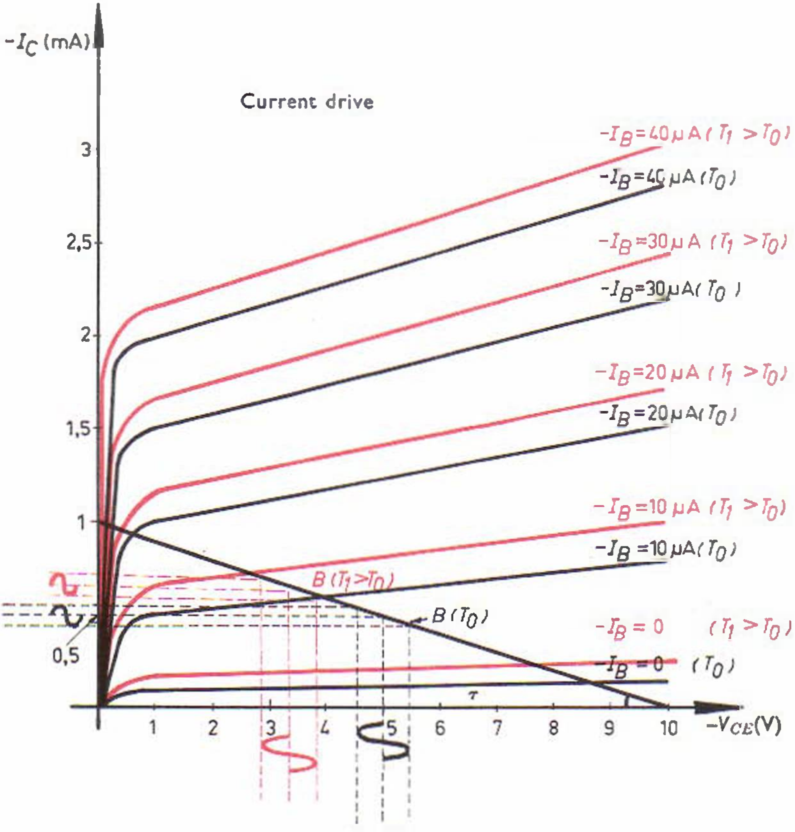

-I_C (mA)
Current drive
3
2,5
2
1,5
1
0,5
-I_B=40 µA (T_1 >T_0)
-I_B=40 µA (T_0)
-I_B=30 µA (T_1 >T_0)
-I_B=30 µA (T_0)
-I_B=20 µA (T_1 >T_0)
-I_B=20 µA (T_0)
-I_B=10 µA (T_1 >T_0)
-I_B=10 µA (T_0)
B (T_1>T_0)
B (T_0)
-I_B = 0 (T_1 >T_0)
-I_B = 0 (T_0)
τ
1
2
3
4
5
6
7
8
9
10
-V_CE(V)

Fig. 270

Current drive

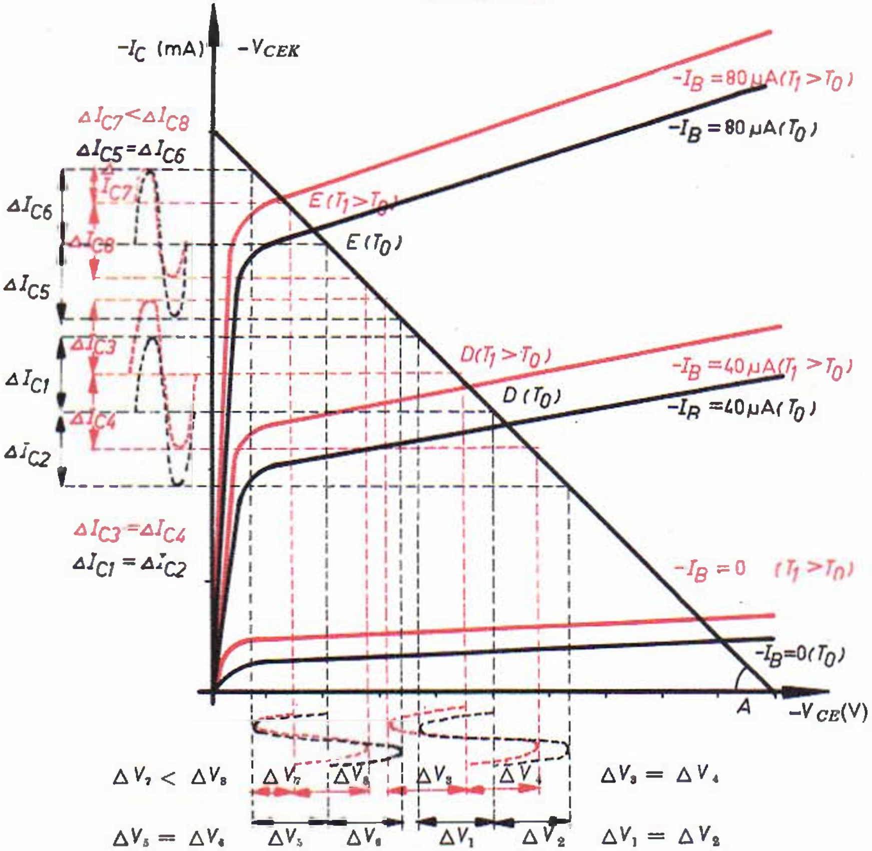


Fig. 271

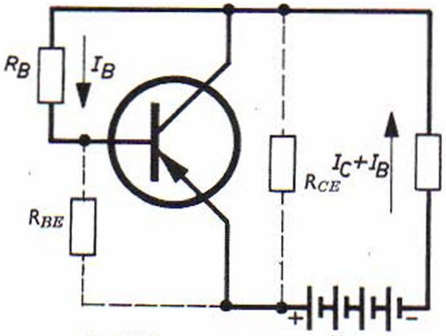

R_B
I_B
R_{BE}
R_{CE}
$I_C + I_B$
R_L
+
–
V_{CC}

Fig. 272

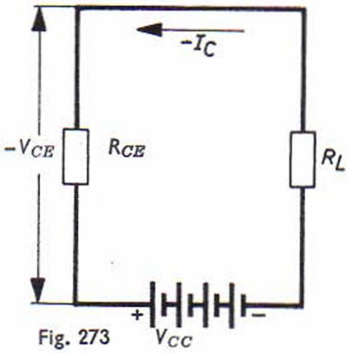

$-I_C$
$-V_{CE}$
R_{CE}
R_L
+
–
V_{CC}

Fig. 273

Large-signal amplifier

The load line marked in the $-I_C = f(-V_{CE})$ characteristic of Fig. 271 is for $R_L = 1\ \text{k}\Omega$. The working point is again chosen so as to be at the centre of the possible range of collector-current and collector-voltage variations. In the figure, the displacement of the working point towards increasing current (or decreasing voltage) is limited by the knee voltage $-V_{CEK}$, and on the decreasing current side by the leakage current $-I_{CEO}$. If the working point selected is close to the $-I_C$ axis under normal conditions (point *E*), an increase in temperature will be accompanied by distortion, as illustrated in Fig. 271. Consequently, when a transistor is employed for large-signal amplification, the working point must be so chosen that the corresponding collector current is considerably lower than the limiting value $-I_{C1}$, or in other words, so that an increase in the leakage current, such as may be caused by a temperature increase, does not give rise to the danger of distortion occurring (point *D* on the characteristic).

27.9. Stabilization by means of a resistor between collector and base

Amongst the various possible ways of stabilizing a transistor against the influence of temperature variations, the simplest solution is to connect a resistor between the collector and the base of the transistor.

Resistive load

Fig. 272 shows a resistor connected between the collector and base, and the transistor also loaded by means of a resistor. The equivalent output circuit is given in Fig. 273. The collector-emitter voltage is equal to the supply voltage minus the voltage drop across the load resistor R_L, due to the passage of the collector current $-I_C$. (The base current $-I_B$ may be neglected in comparison with the collector current $-I_C$). Therefore:

$$-V_{CE} = -V_{CC} - (-R_L I_C).$$

The equivalent base-emitter circuit is given in Fig. 274. The base-emitter voltage is equal to the difference between the collector-emitter voltage and the voltage drop across the resistor R_B caused by the base current $-I_B$:

$$-V_{BE} = -V_{CE} - (-R_B I_B).$$

For a given temperature we can determine the collector-emitter voltage, the collector current, the base-emitter voltage and the base current, the last of which is dependent on the base-emitter voltage and thus also on the collector-emitter voltage. An increase in temperature will cause the collector current to increase, and this means that the collector-emitter voltage will decrease, since $-V_{CC}$ remains unchanged.

If $-V_{CE}$ and therefore $-V_{CB}$ decreases, $-R_B I_B$ and therefore $-I_B$ will also decrease. We can select the value of R_B such that the change in $-R_B I_B$ is small in comparison with $-V_{CE}$. An increase in temperature is then accompanied by a slight decrease in the base current, and since this is multiplied by the gain h_{FE} of the transistor, the decrease of $-h_{FE} I_B$ will be sufficient to cancel out the increase in $-I_C$.

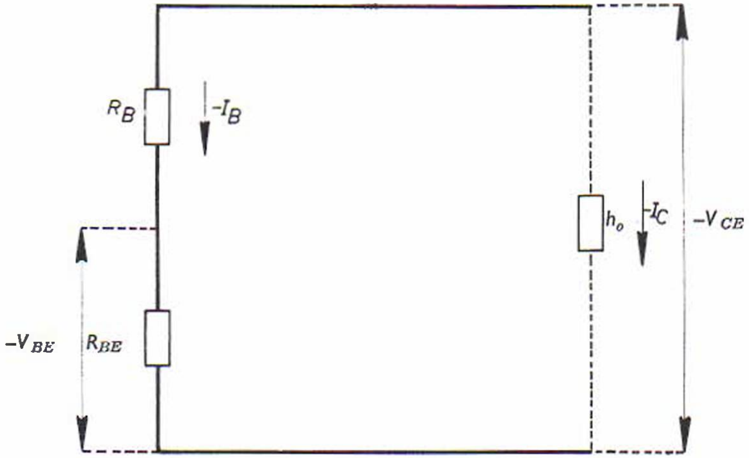

R_B
$-I_B$
$-V_{BE}$
R_{BE}
h_o
$-I_C$
$-V_{CE}$

Fig. 274

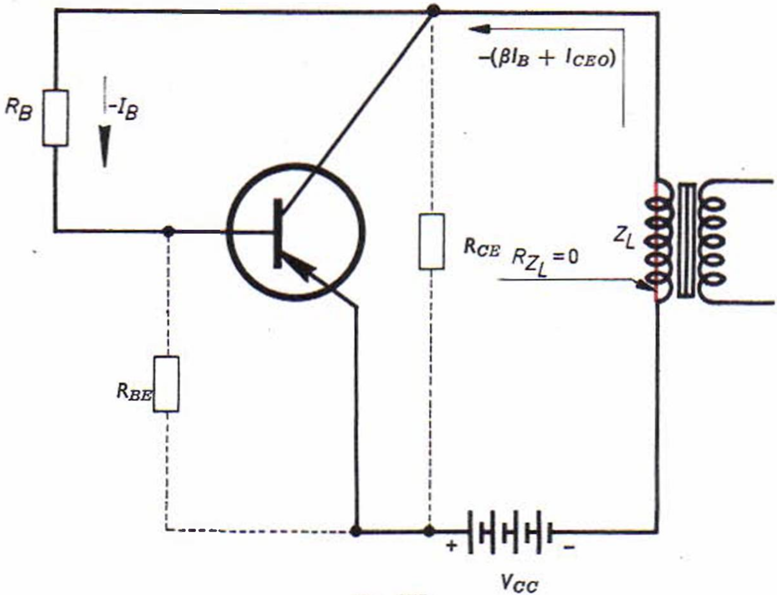

R_B
$-I_B$
$-(\beta I_B + I_{CEO})$
Z_L
R_{CE}
$R_{Z_L} = 0$
R_{BE}
+
-
V_{CC}

Fig. 275

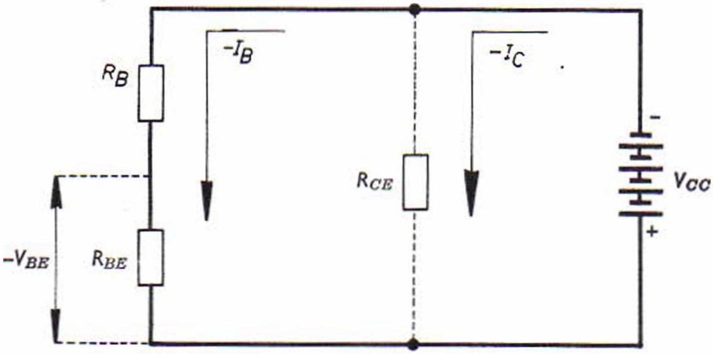

R_B
$-I_B$
$-I_C$
R_{CE}
-
V_{CC}
+
$-V_{BE}$
R_{BE}

Fig. 276

Load consisting of a transformer

In the circuit of Fig. 275 the transistor is loaded by a transformer whose primary impedance is Z_L. Since the resistance of the primary of this transformer may be neglected, the collector-emitter voltage $-V_{CE}$ is taken as equal to the supply voltage $-V_{CC}$.

Fig. 276 represents the equivalent input circuit of the transistor. This consists of a direct voltage source ($-V_{CC} = -V_{CE}$) in series with which is connected the input impedance of the transistor, and the stabilizing resistor R_B. The base-emitter voltage is equal to the supply voltage minus the voltage drop produced by the base current $-I_B$ across the resistor R_B:

$$-V_{BE} = -V_{CC} - (-R_B I_B).$$

When the temperature rises the collector current will increase but the base-emitter voltage and thus the base current remain unchanged. This is because $-V_{CC}$ remains unchanged, and therefore $-R_B I_B$ as well. Consequently the rise in collector current is not opposed in any way.

The collector dissipation is given by the equation:

$$P_C = V_{CE} I_C$$

in which I_C increases with the temperature, while $V_{CE} = V_{CC}$ remains constant, so that P_C also increases with temperature. Consequently the connection of a resistor between the collector and base of a transistor with transformer load does not produce any improvement whatever in the thermal stability of the transistor.

It is thus desirable to have a circuit by means of which the transistor can be stabilized under all conditions. Such a circuit, in which a resistor is connected in the emitter circuit, is described below. However, a resistor connected between the collector and base of a transistor can be useful for stabilization against temperature variations in those cases in which the load consists of the coil of a loudspeaker, the resistance of which does not differ much from its impedance.

27.10. Stabilization by a resistor connected in series with the emitter

A resistor in series with the emitter can ensure good stabilization against temperature variations for both resistive and inductive loads.

Resistive load

In the circuit of Fig. 277 the transistor is loaded by a resistor R_L. Let us suppose that the supply voltage of the collector-emitter circuit is $-V_{CC}$, and that the base-emitter bias is provided by a battery of voltage $-V_{BB}$. Fig. 278 shows the equivalent output circuit. This contains a direct voltage source in series with which are connected the load resistor R_L, the output resistance h_o of the transistor, and the emitter resistor R_E. Through this circuit flows the collector current $-I_C$.

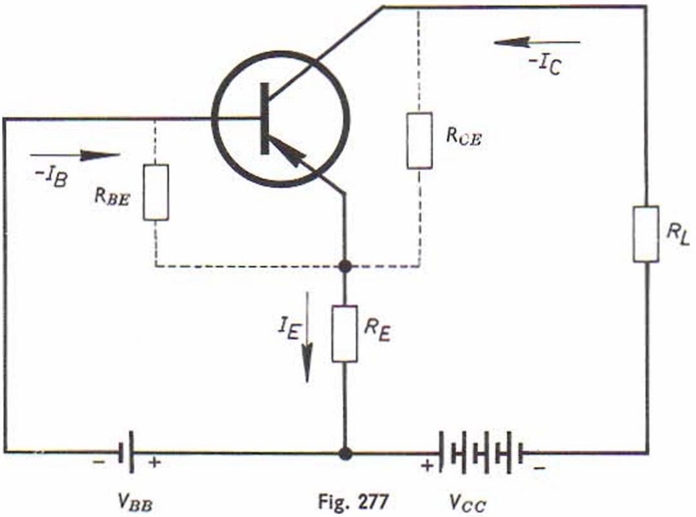

-I_C
R_{CE}
-I_B
R_{BE}
R_L
I_E
R_E
–
+
V_{BB}
+
–
V_{CC}

Fig. 277

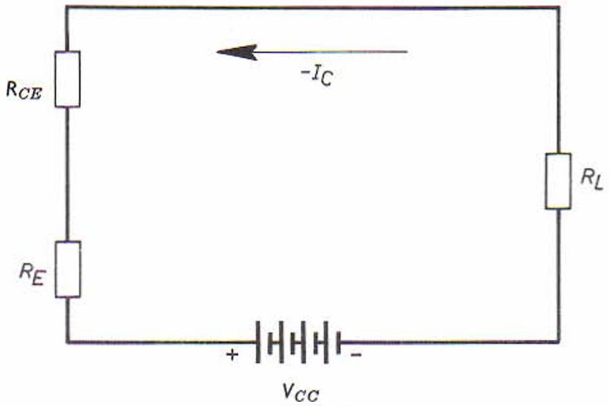

-I_C
R_{CE}
R_L
R_E
+
–
V_{CC}

Fig. 278

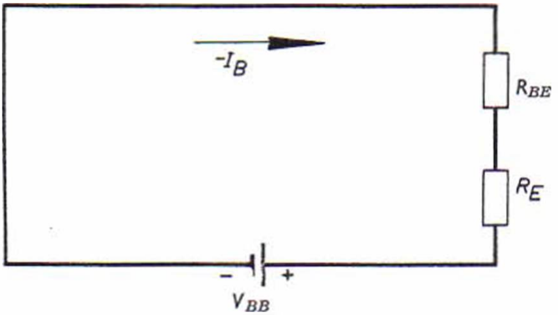

$-I_B$
R_{BE}
R_E
-
+
V_{BB}

Fig. 279

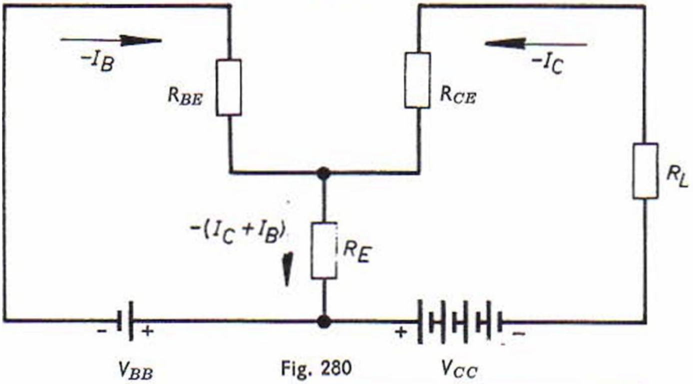

$-I_B$
R_{BE}
R_{CE}
$-I_C$
R_L
$-(I_C+I_B)$
R_E
-
+
V_{BB}
+
-
V_{CC}

Fig. 280

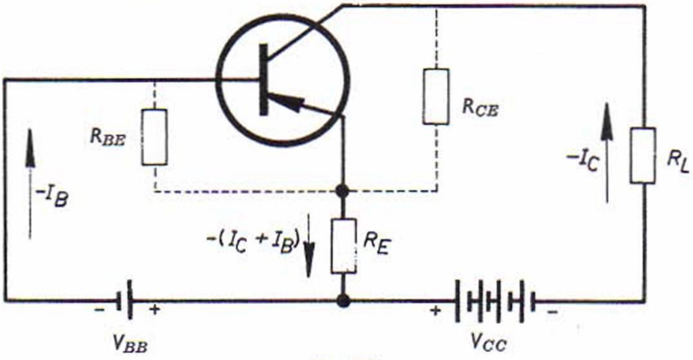

R_{BE}
R_{CE}
$-I_B$
$-I_C$
R_L
$-(I_C+I_B)$
R_E
-
+
V_{BB}
+
-
V_{CC}

Fig. 281

The equivalent input circuit is represented in Fig. 279. This consists of a direct voltage source $-V_{BB}$, in series with which are connected the input resistance R_i of the transistor and the emitter resistor R_E. Through this circuit flows the base current $-I_B$. If we combine the equivalent input and output circuits (Fig. 280), we see that both circuits have the resistor R_E in common, so that both the collector and base currents flow through this resistor.

The base-emitter voltage, i.e. the voltage across the input impedance, is equal to the bias voltage $-V_{BB}$ minus the voltage drop across the resistor R_E (see Fig. 281). This voltage drop is given by the value of R_E multiplied by the current flowing through the resistor, that is, by the emitter current of the transistor:

This current is:

$$-I_E = -I_C - I_B,$$

so that the voltage drop across this resistor equals:

$$-V_{R_E} = R_E \,.\, (-I_C - I_B).$$

Now $-I_B$ may be neglected in comparison with $-I_C$ and the voltage drop across R_E may be equated to $R_E\ I_C$. Consequently, the base-emitter voltage is:

$$-V_{BE} = -V_{BB} + R_E I_C.$$

If the temperature increases, the collector current will also increase. This means that the second term in the above equation will also increase, and since $-V_{BB}$ remains constant, $-V_{BE}$ will decrease. The base current $-I_B$ decreases as a result of this reduction in $-V_{BE}$.

The collector current $-I_C$ is given by:

$$-I_C = -(h_{FE}I_B + I_{CEO}).$$

The decrease of $h_{FE}I_B$ opposes the increase in $-I_C$ so that the change in the collector dissipation resulting from temperature variations will decrease. This means that the transistor is stabilized against temperature variations.

Inductive load

Fig. 282 represents the case when the load is formed by a self inductance Z_L, the primary of a transformer, the resistance of which may be neglected. The equivalent input and output circuit is shown in Fig. 283. The base-emitter voltage is again equal to the bias voltage $-V_{BB}$ minus the voltage-drop across the emitter resistor R_E. In this case also, this voltage drop is determined by the value of R_E and of the emitter current $-I_E$, which equals the sum of the collector current $-I_C$ and the base current $-I_B$. Since $-I_B$ may be neglected in comparison with $-I_C$ the voltage-drop across the emitter resistor is thus:

$$-V_{R_E} = -R_E I_C$$

while the base-emitter voltage is equal to:

$$-V_{BE} = -V_{BB} + V_{R_E} = -V_{BB} + R_E I_C.$$

Here too, a temperature rise will cause the collector current to increase, with the result that $R_E I_C$ will also increase, and since $-V_{BB}$ remains constant, $-V_{BE}$ will decrease. The result of this is that the base current $-I_B$ also decreases.

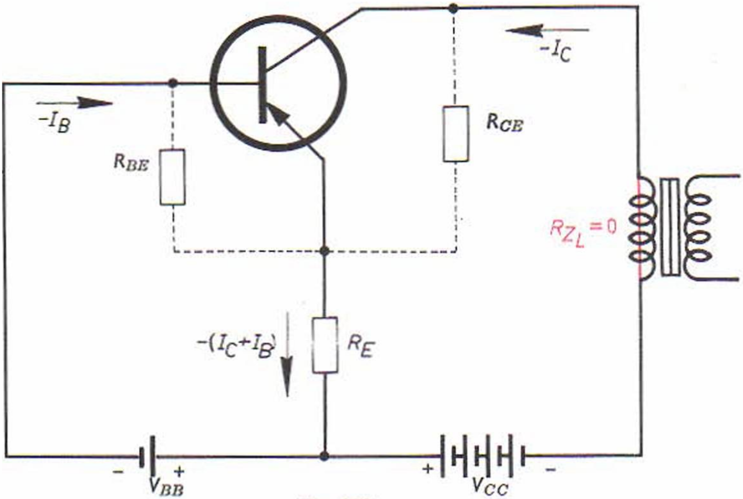

-I_C
-I_B
R_{BE}
R_{CE}
$R_{Z_L}=0$
-(I_C+I_B)
R_E
-
+
V_{BB}
+
V_{CC}
-

Fig. 282

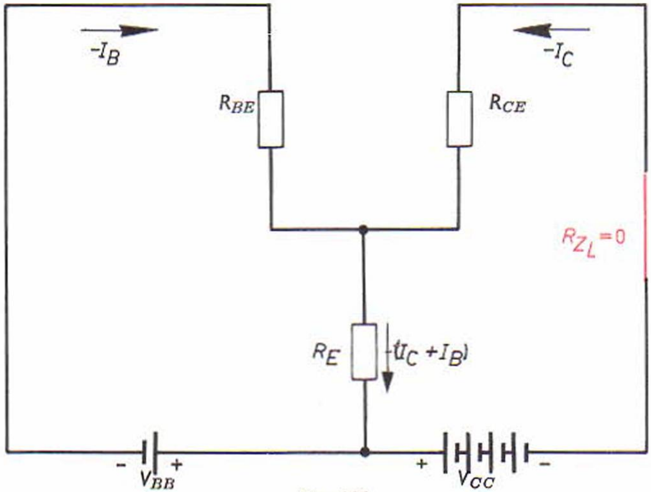
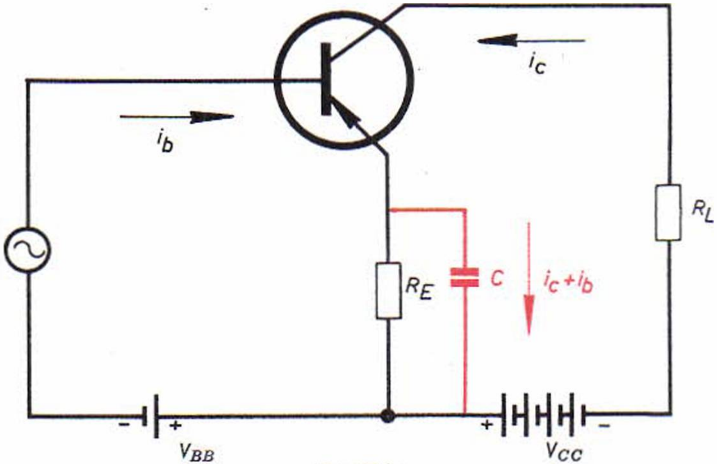
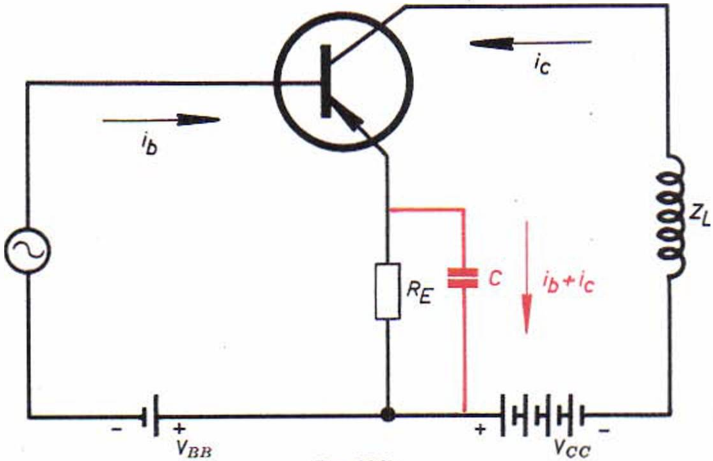

-I_B
-I_C
R_{BE}
R_{CE}
$R_{Z_L}=0$
R_E
-(I_C +I_B)
-
+
V_{BB}
+
V_{CC}
-

Fig. 283



i_c
i_b
R_L
R_E
C
$i_c + i_b$
– +
V_{BB}
+ –
V_{CC}

Fig. 284a



i_c
i_b
Z_L
R_E
C
$i_b + i_c$
– +
V_{BB}
+ –
V_{CC}

Fig. 284b

Consequently, the collector current $-I_C$ will remain practically constant, since the decrease in $-h_{FE}I_B$ opposes the increase in $-I_C$. The collector dissipation $P_C = V_{CE}I_C$ thus varies only little with the temperature. As in the previous case, the transistor is stabilized against temperature variations.

This circuit is based on the principle of negative d.c. feedback but this feedback also acts on the a.c. signal which is to be amplified. If such a signal is conveyed to the input of the amplifier, a voltage $v_e = R_E i_e$ will arise across the emitter resistor. This voltage is in phase with the input voltage, giving rise to a negative feedback which reduces the amplification of the circuit. In order to oppose this effect, the emitter resistor is shunted by means of a capacitor C (see Figs. 284a and b). The object of this capacitor is to keep the voltage across the emitter resistor constant, i.e. to nullify the a.c. feedback.

We must point out that this method of stabilization will only work effectively if the bias voltage $-V_{BB}$ is constant. It is easy to satisfy this condition by drawing the bias voltage from a battery, but there are numerous objections against doing this, so that it would be desirable to obtain the same result by taking the bias voltage from the supply battery $-V_{CC}$. We shall show that this is indeed possible, by connecting the base to a voltage divider, in combination with a series resistor in the emitter lead.

27.11. Base connected to a voltage divider, combined with the use of an emitter resistor

In the circuit of Fig. 285, the required base bias is obtained by means of a voltage divider, while stabilization against variations of temperature is achieved by the inclusion of an emitter resistor. The transistor is loaded by means of a resistance or inductance. The equivalent input circuit is shown in Fig. 286. The resistor R_1 is connected in series with resistor R_2 across a direct voltage source $-V_{CC}$, while the input resistance R_i is connected in series with the emitter resistor R_E, which is in parallel with resistor R_2. The voltage V_{R2} across the terminals of resistor R_2 is equal to the bias voltage $-V_{BB}$.

The equivalent circuit diagram shows that the current through R_1 is equal to the sum of the potential divider current I_p and the base current I_B. The potential divides current I_p flows through resistor R_2, the base current I_B flows through the input resistance R_i, and the current flowing through the emitter resistor R_E is equal to the sum of the collector current I_C and the base current I_B which may be neglected here. Fig. 287 shows the equivalent circuit diagram of the bias and stabilization circuits of the transistor. The bias voltage $-V_{BB}$ is given by the voltage drop across the resistor R_2:

$$-V_{R2} = -V_{BB} = -V_{CC} + R_1(I_p + I_B).$$

This bias voltage must be constant; it is equal to the supply voltage minus the voltage drop across resistor R_1, through which flow both I_p and I_B. Now $-V_{CC}$ is constant, while $R_1(I_p + I_B)$ will also be practically constant provided that I_p is very much greater than I_B. In this case, I_B may be neglected in comparison with I_p, and the bias voltage will be equal to:

$$-V_{BB} = -V_{CC} + R_1 I_p.$$

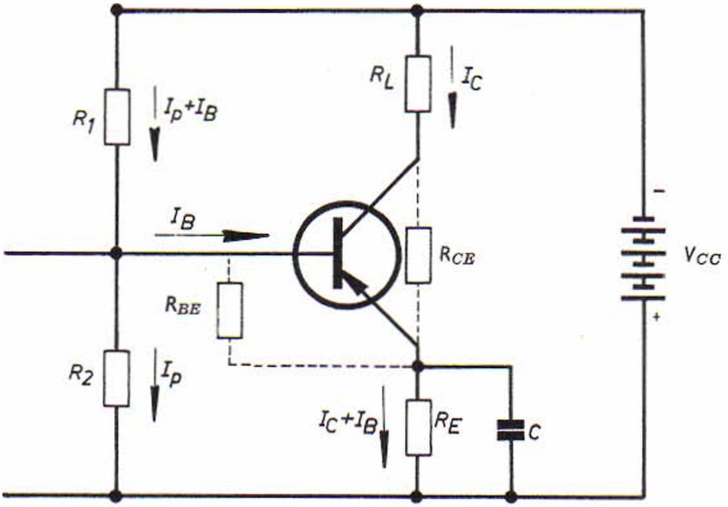

R_1
I_p+I_B
R_L
I_C
I_B
R_{CE}
V_{CC}
R_{BE}
R_2
I_p
I_C+I_B
R_E
C
-
+

Fig. 285

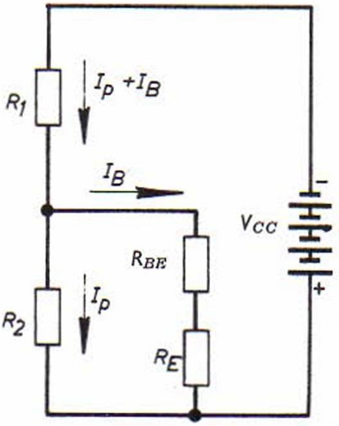

R_1
I_p+I_B
I_B
V_{CC}
R_{BE}
R_2
I_p
R_E
-
+

Fig. 286

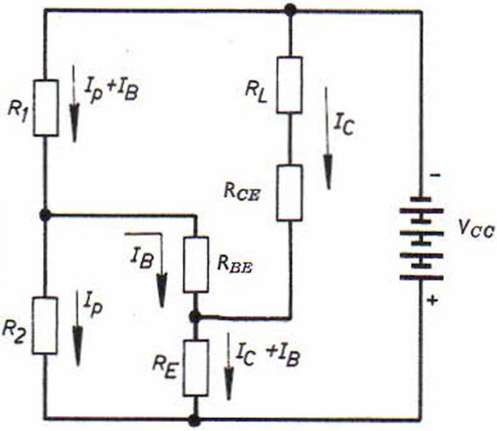

R_1
I_p+I_B
R_L
I_C
R_{CE}
V_{CC}
I_B
R_{BE}
R_2
I_p
R_E
I_C+I_B
-
+

Fig. 287

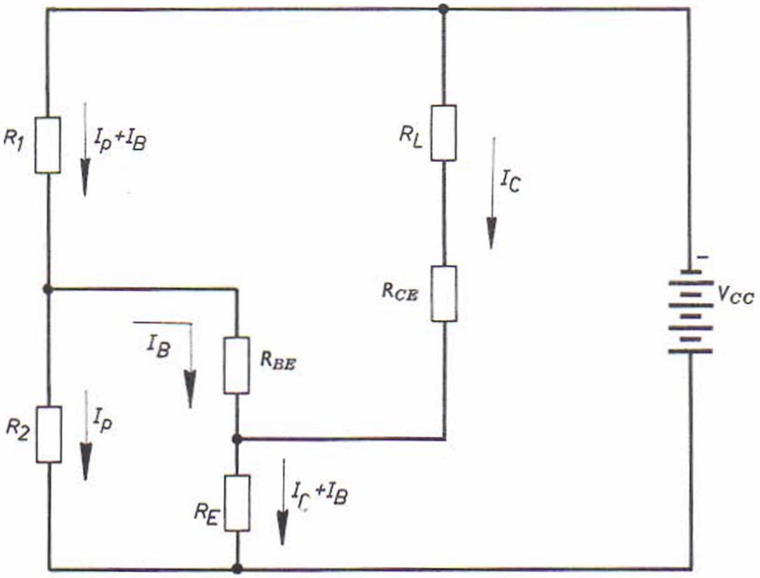

R_1
I_p+I_B
R_L
I_C
R_{CE}
V_{CC}
I_B
R_{BE}
R_2
I_p
R_E
I_C+I_B

Fig. 288

The base-emitter voltage is given by:

$$-V_{BE} = -V_{BB} + R_E (I_C + I_B)$$

since, as shown in Fig. 288, both the collector current and the base current flow through the emitter resistor. As we can neglect I_B in relation to I_C we may write the base-emitter voltage as:

$$-V_{BE} = -V_{BB} + R_E I_C = -V_{R_2} + R_E I_C.$$

Since it was assumed that the ballast current I_p is much larger than the base current, V_{R_2} may be taken as constant. Now an increase in the ambient temperature will mean that the collector current

$$-I_C = -\mathrm{h}_{FE} I_B - I_{CEO}$$

will increase. The result of this increase is that the second term in the equation for $-V_{BE}$ will also increase, so that this voltage decreases.

A decrease in $-V_{BE}$ means that the base current $-I_B$ will also decrease, so that the increase in $-I_C$ is compensated by the decrease in $-\mathrm{h}_{FE} I_B$. In this way, the collector current $-I_C$ and also the collector dissipation P_C will remain practically constant.

As we have already mentioned, care must be taken to see that the bias voltage, i.e. the voltage drop across the resistor between the base of the transistor and the positive pole of the battery, remains constant under all conditions. To achieve this, the ballast current through the voltage divider should be much greater than the base current. In theory, with a base current of 100 μA, the current flowing through the voltage divider could be 1000 times as large i.e. a current of 100 mA. Now one of the most important advantages of transistor amplifiers is their low power consumption, and if each transistor were to require a voltage divider with such a large current flowing through it, this advantage would be completely cancelled out. In practice therefore, one has to accept a much lower current, usually from five to ten times the size of the base current. This method of stabilization is extremely satisfactory, independent of whether the transistor load is resistive or inductive. A closer examination of the output circuit of the transistor (Fig. 290), shows that the supply voltage is divided into three portions.:

> The voltage across the load impedance R_L in the collector circuit.
> The voltage across the output resistance h_O of the transistor. (collector-emitter voltage $-V_{CE}$).
> The voltage across the emitter resistor R_E.

The collector current in a transistor in a pre-amplifier stage will be of the order of a few mA so that the resulting voltage drop across resistor R_E will be very low. In this case it is possible to use fairly high resistances, of the order of a few hundred ohms.

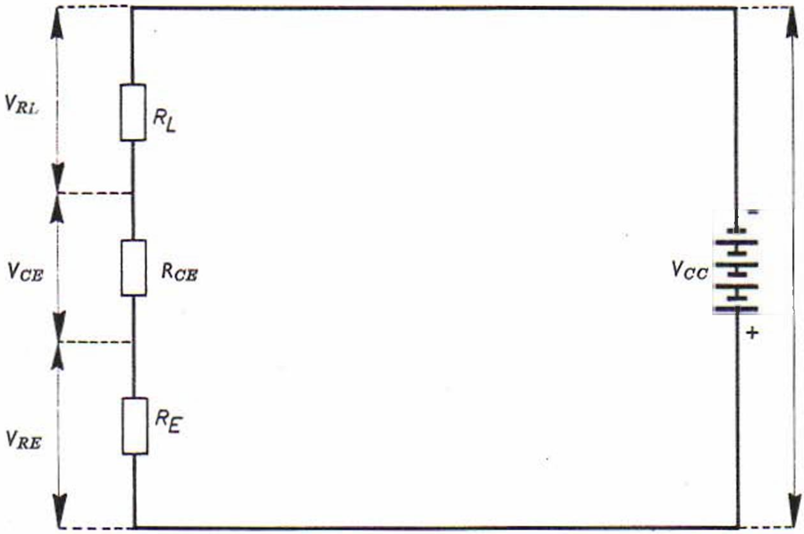

V_{RL}
R_L
V_{CE}
R_{CE}
V_{RE}
R_E
V_{CC}
-
+

Fig. 290

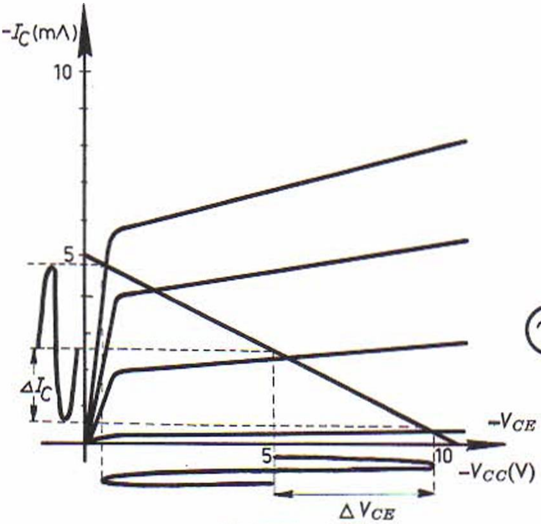


Fig. 291

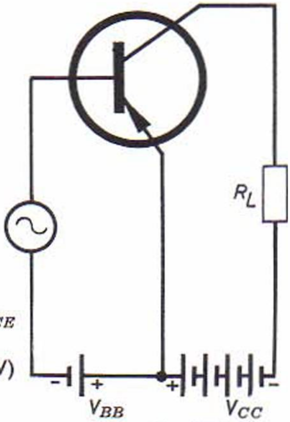


Fig. 292

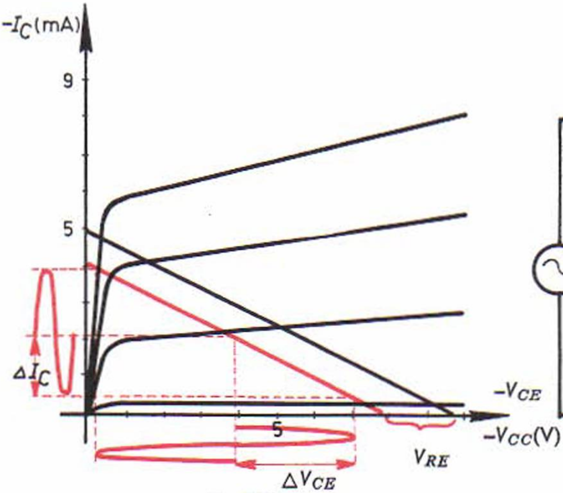


Fig. 293

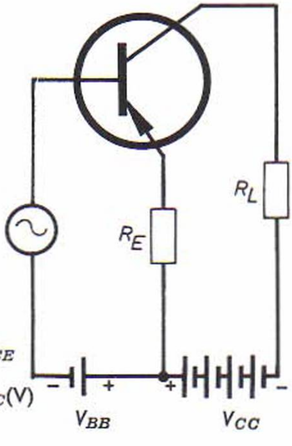

R_E
R_L
–
+
V_{BB}
+
–
V_{CC}

Fig. 294

The maximum output of the transistor depends on the r.m.s. values of the collector current and the collector-emitter voltage:

$$P_o = (\Delta V_{CE}/\sqrt{2})\,(\Delta I_C/\sqrt{2}).$$

In the characteristic of Fig. 291 the peak-to-peak value of the collector-emitter voltage, in the absence of a series resistor in the emitter circuit (Fig. 292), is equal to the supply voltage $-V_{CC}$ (minus the knee voltage $-V_{CEK}$, which we can neglect in this case).

The peak-to-peak value of the collector-emitter voltage is thus practically equal to the supply voltage:

$$2\Delta V_{CE} = -V_{CC}.$$

If a resistor (Fig. 294) is now connected in series with the emitter, the family of characteristics in Fig. 293 shows that the peak-to-peak value of the collector-emitter voltage is equal to the supply voltage $-V_{CC}$ minus the voltage drop across the emitter resistor:

$$2\Delta V_{CE} = -V_{CC} + R_E I_C.$$

The resistor included in the emitter circuit is thus found to reduce the peak-to-peak value of the collector-emitter voltage, or in other words the amplitude of the displacement of the working point along the load line decreases, and with it the maximum output power of the transistor.
For the amplification of higher power, however, it will usually be desirable for the transistor to deliver maximum output, so that in this case it is desirable to have another method of stabilization available or to select a higher battery voltage. A frequently-used method of stabilization is based on the characteristic of a thermistor, the resistance of which decreases as the temperature increases.

27.12. Voltage divider with thermistor

Fig. 295 shows a circuit in which the bias voltage is derived from a voltage-divider consisting of a linear resistance R_1 in series with a thermistor R_{th}. In this circuit the emitter is connected directly to the positive pole of the battery.

Fig. 296 shows the equivalent circuit diagram of the input circuit. Through the resistor R_1 flows the potential divider current I_p and the base current I_B. Through the thermistor flows the potential divider current, and through the input resistance of the transistor flows the transistor base current. The base-emitter voltage, i.e. the voltage across the input resistance of the transistor, is equal to the voltage drop across the thermistor:

$$-V_{R_{\text{th}}} = -V_{BE} = -V_{CC} + R_1 (I_p + I_B).$$

If the potential divider current is much greater than the base current the latter may be neglected, so that the base-emitter voltage is then:

$$-V_{BE} = -V_{CC} + R_1 I_p = R_{\text{var}} I_p.$$

If the ambient temperature rises, the current $-I_C$ will increase. Since the resistance of the thermistor decreases as the temperature increases, a rise in temperature will mean that the voltage drop across the thermistor, that is, the base-emitter voltage of the transistor, will also decrease.

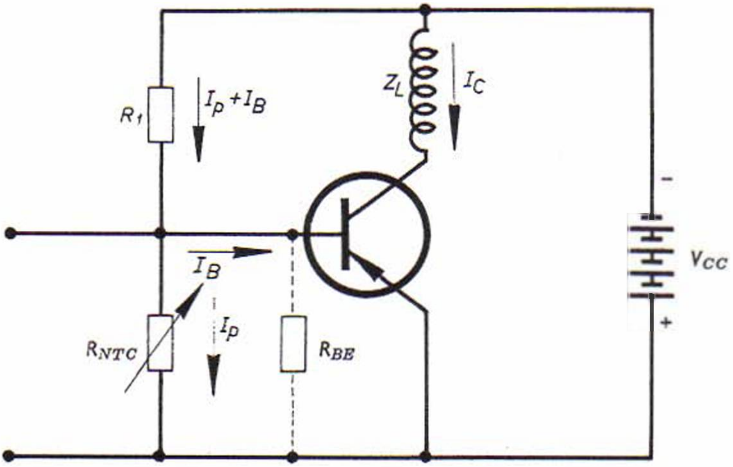

R_1
$I_p + I_B$
Z_L
I_C
-
V_{CC}
+
I_B
R_{NTC}
I_p
R_{BE}

Fig. 295

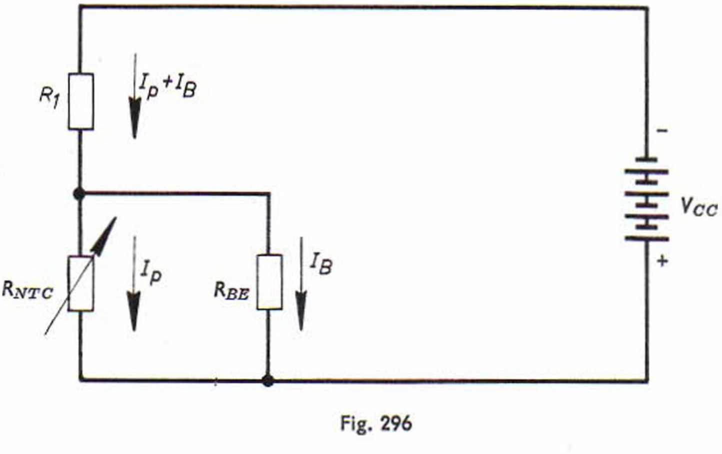

R_1
$I_p + I_B$
-
V_{CC}
+
R_{NTC}
I_p
R_{BE}
I_B

Fig. 296

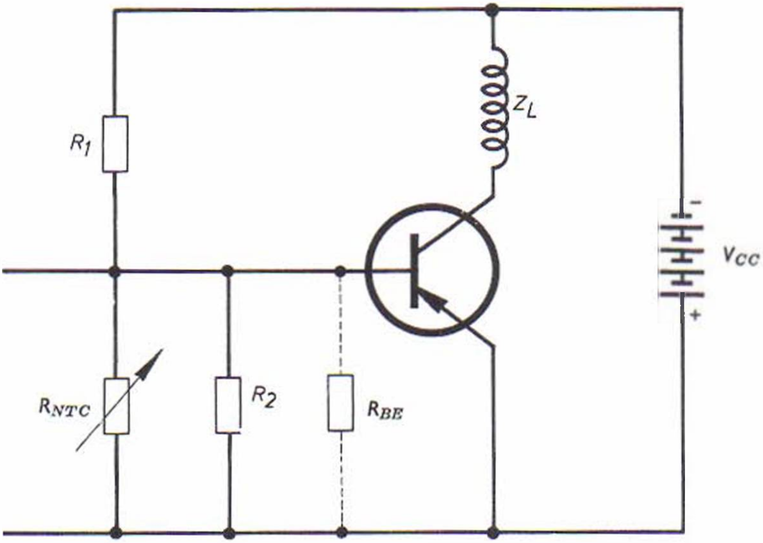

R_1
Z_L
V_{CC}
-
+
R_{NTC}
R_2
R_{BE}

Fig. 297

If the base-emitter voltage decreases, the base current $-I_B$ also decreases. In this way the increase in $-I_C$ resulting from an increase in temperature is opposed by the decrease in $-I_B$. The influence of temperature variations is thus compensated and it is even possible to obtain over-compensation. In this circuit the whole supply voltage $-V_{CC}$ is utilized, either in the output circuit or across the transistor terminals; the elimination of the resistor in series with the emitter means that the circuit can yield the maximum output power. It is also possible to employ a combination of the last two methods.

Fig. 297 corresponds in broad outline to a combination of the circuits of Figs. 285 and 295. The voltage divider is formed here by the resistor R_1 and the resistor R_2 with the thermistor R_{th} connected in parallel with it. The operation of this circuit is very similar to that of the circuits already discussed; however, the base-emitter voltage depends on the mutual relationship between the values of R_2 and R_{th}, so that the stabilizing effect of the thermistor is not so great.

Whichever circuit is employed, it is always necessary to stabilize the transistor against the effect of temperature variations. The ambient temperature is the fundamental quantity controlling this compensation; we shall see that under certain conditions the ambient temperature has to be taken into account even in spite of the stabilization that is applied. At normal ambient temperatures however, it is always possible, with the aid of the circuits just described, to prevent the transistor being overheated or damaged.

CHAPTER 28

The load impedance

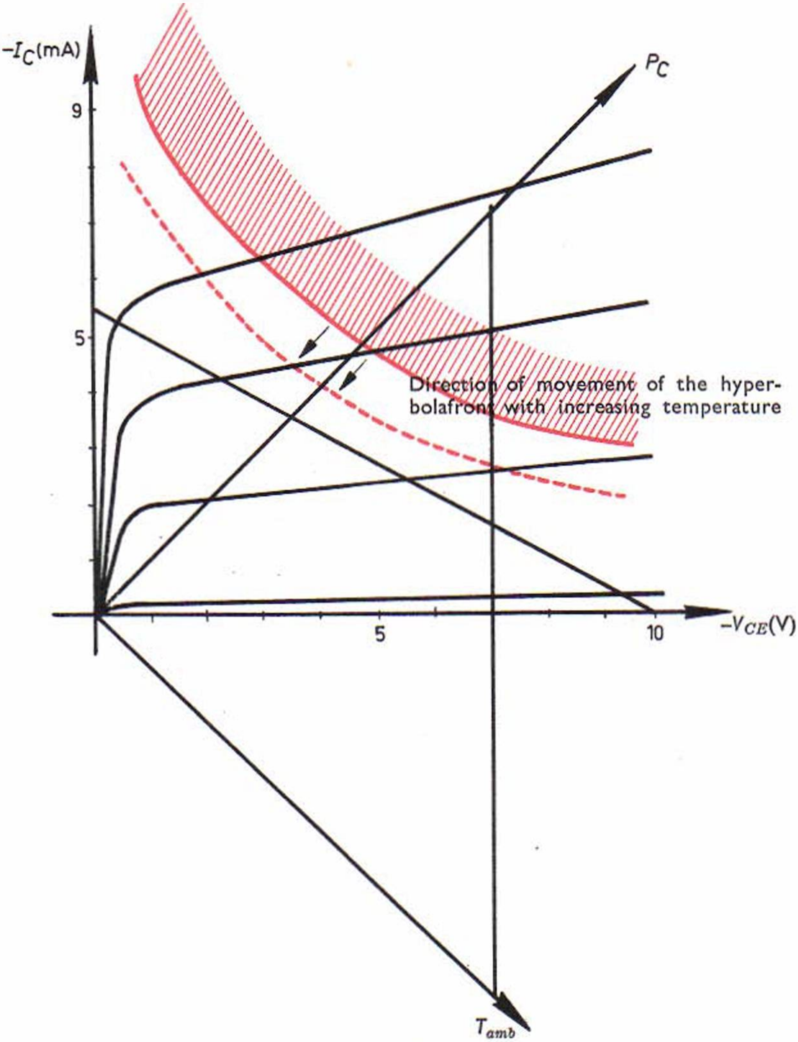
The operation of an electron tube can be investigated by drawing the load line on the $I_a = f(V_a)$ characteristic. In a similar way the operation of a transistor as amplifier can be studied with the aid of the load line in the $-I_C = f(-V_{CE})$ characteristic. In doing this, the following four important factors must be taken into account:

The equal power hyperbola.
The maximum output power of the amplifier.
The choice of the constant collector current (bias voltage of the transistor).
The maximum power gain.

28.1. Constant-power hyperbola

This factor has already been considered on page 170. Fig. 298 represents the $-I_C = f(-V_{CE})$ characteristic of the transistor. On this figure we can plot the hyperbola which represents the locus of the maximum output power for a given temperature. At no point may the load line come within the hatched area to the right-hand side of this hyperbola, as all points in this area correspond to a greater power than the maximum permissible collector dissipation at the temperature in question.

In practice, the load impedance must always be chosen so that the load line remains at some distance from the hyperbola; this is because it is very difficult to foresee the values which the ambient temperature may reach, and an increase in this temperature is expressed in a displacement of the hyperbola to the left, as indicated by the arrows in the figure.



$-I_C$(mA)
9
5
P_C
Direction of movement of the hyper-
bolafront with increasing temperature
5
10
$-V_{CE}$(V)
T_{amb}

Fig. 298

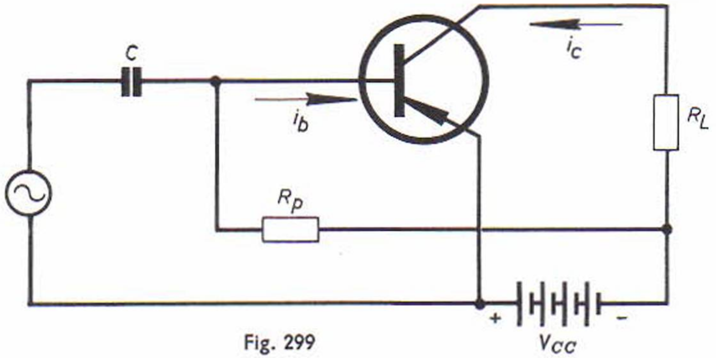

C
i_b
i_c
R_L
R_p
+
−
V_{CC}

Fig. 299

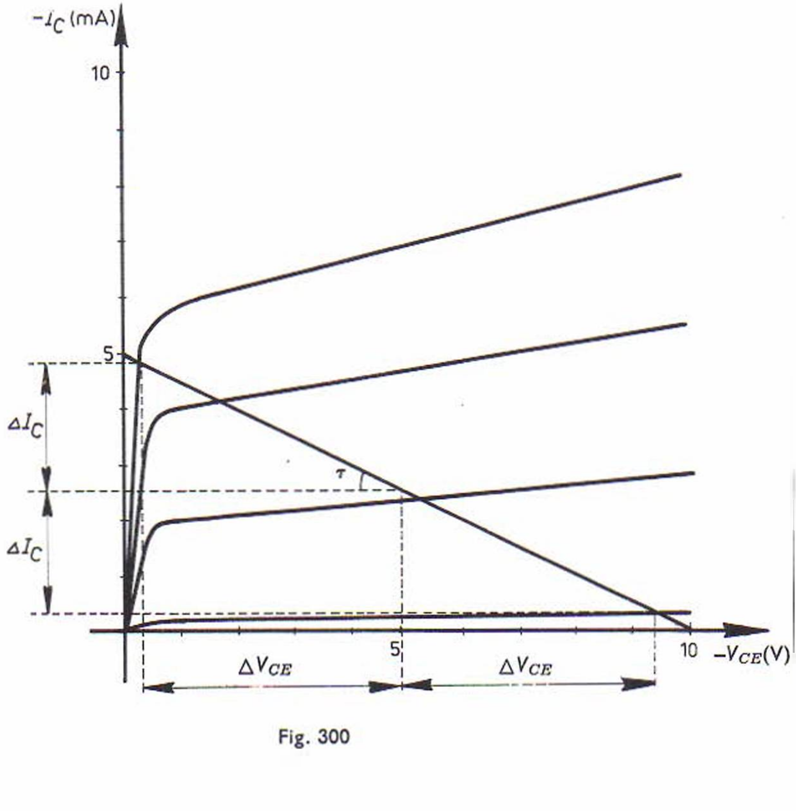


Fig. 300

28.2. Maximum output power of an amplifier

In order to determine the maximum output power of an amplifier, we must distinguish between the following two cases:

Resistive load.
Inductive load.

Resistive load

Fig. 299 shows the amplifier loaded by a resistance R_L. In the characteristic of Fig. 300, we determine the point corresponding to $-V_{CC} = 10$ V. In order to obtain the maximum output power with minimum distortion, the voltage must be able to swing the same distance on both sides of the working point. This symmetrical displacement must correspond to the maximum amplitude. The working point is now given by the equation:

$$-V_{CE} = -V_{CC}/2 = -10/2 = -5 \text{ V}.$$

The power available at the output equals the product of the maximum variation of the collector-emitter voltage and the maximum variation of the collector current:

$$P_o = (\Delta V_{CE}/\sqrt{2})\,(\Delta I_C/\sqrt{2}).$$

There are both top and bottom limits to the load resistance (R_L).

Low value of R_L

The tangent of the angle between the load line and the horizontal axis is given by $1/R_L$, so that this tangent – i.e. the slope of this line – increases as the value of R_L decreases.

The constant-power hyperbola sets a limit to the slope of the load line, as the latter must not come within the hatched region in Fig. 301, which means that R_L must not be less than 1 kΩ. This value corresponds to the maximum output power.

In the absence of an input signal the collector-emitter voltage must be 5 V, corresponding to point A on the load line. The collector current is then $-I_C = 5$ mA and ΔV_{CE} and ΔI_C are practically equal to 5 V and 5 mA respectively. The maximum power available at the collector is then equal to:

$$P_o = (\Delta V_{CE}/\sqrt{2}).\ (\Delta I_C/\sqrt{2}) = (4.4 \times 5/2)10^{-3} = 11 \times 10^{-3}\ \text{W} = 11\ \text{mW}.$$

High value of R_L

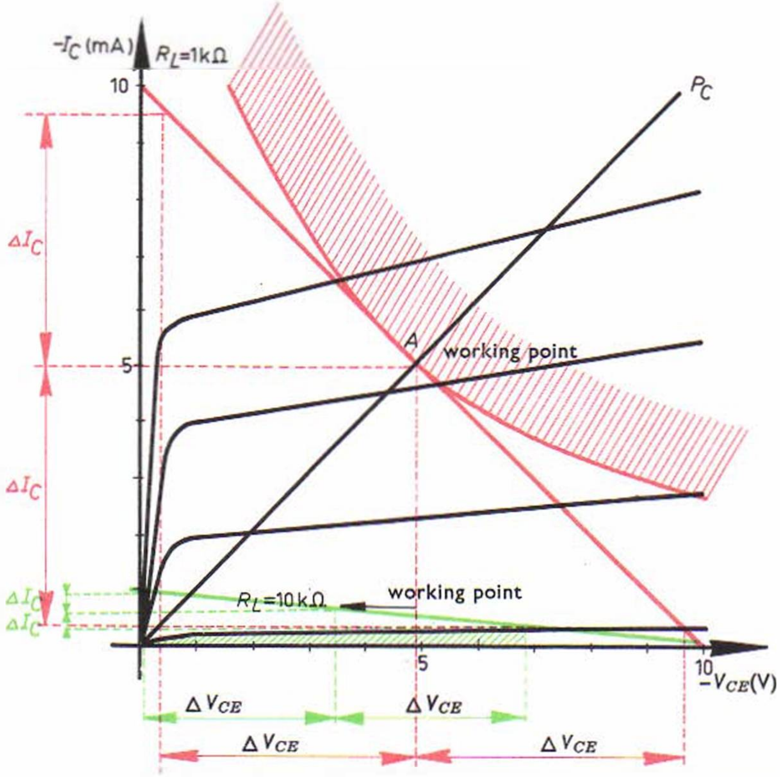
If R_L has a high value the load line will only have a slight slope. In this case too, the operating conditions of the transistor will be chosen so that the voltage can move an equal distance to both sides of the working point, for the sake of obtaining minimum distortion.

Consequently the value of the constant collector current will decrease with decreasing slope of the load line, i.e. with increasing value of R_L.

Suppose that $R_L = 10$ kΩ. The corresponding load line is printed green in Fig. 301. If the transistor operating conditions are now selected so that the working point corresponds to $-V_{CE} = 3.6$ V in the absence of an input signal, the constant collector current will only equal 0.65 mA. In this case, ΔV_{CE} and ΔI_C equal approximately 3.2 V and 0.4 mA respectively, so that the maximum power available at the output equals:

$$P_o = (\Delta V_{CE}/\sqrt{2}).\ (\Delta I_C/\sqrt{2}) = (3.2 \times 0.4/2)10^{-3} = 0.64 \times 10^{-3} = 0.64\ \text{mW}.$$

We thus see that the maximum power available at the collector decreases as the load impedance increases.



$-I_C$ (mA)
R_L=1kΩ
10
5
P_C
A
working point
R_L=10kΩ
working point
ΔI_C
ΔI_C
ΔI_C
ΔI_C
5
10
$-V_{CE}$(V)
ΔV_{CE}
ΔV_{CE}
ΔV_{CE}
ΔV_{CE}

Fig. 301

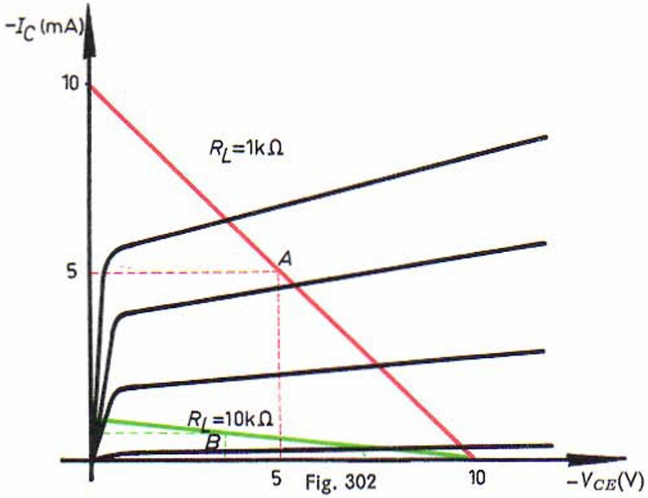


Fig. 302

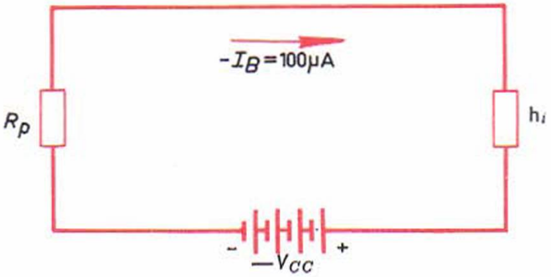

$-I_B$=100µA
R_p
h_i
-
+
$-V_{CC}$

Fig. 303

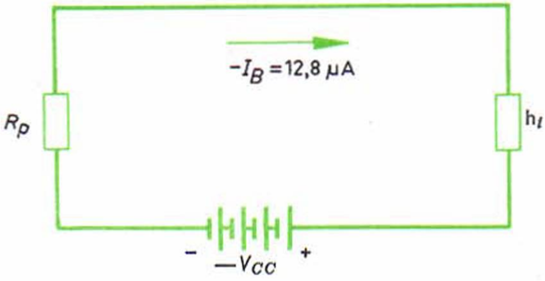

$-I_B$=12,8 µA
R_p
h_i
-
+
$-V_{CC}$

Fig. 304

The effect of the load impedance on the transistor bias voltage

As the relationship between the constant collector current and the constant base current is given by the equation $-I_C = -h_{FE}I_B$, the constant base current will decrease with increasing value of the load impedance R_L. Suppose that h_{FE}, the current amplification factor of the transistor, is equal to 50.

At a low value of R_L, e.g. 1 kΩ (red line in Fig. 302) we have $-I_C = 5$ mA so that, neglecting $-I_{CEO}$ in relation to $-I_C$:

$$-I_B = -I_C/h_{FE} = 5/50 = 0.1 \text{ mA} = 100\ \mu\text{A}.$$

The equivalent input circuit is shown in Fig. 303. The total resistance of the circuit can be calculated from the supply voltage $-V_{CC}$ and the base current $-I_B$:

$$R_p + h_i = -V_{CC}/-I_B = 10/10^{-4} = 10^5\Omega = 100 \text{ k}\Omega.$$

Since the input resistance h_i can be neglected in comparison with the series resistor R_p, the latter can be taken as 100 kΩ. At a high value of R_L, e.g. 10 kΩ (green line), we have $-I_C = 0.5$ mA, so that, neglecting $-I_{CEO}$ in relation to $-I_C$:

$$-I_B = -I_C/h_{FE} = 0.64/50 = 0.0128 \text{ mA} = 12.8\ \mu\text{A}.$$

The equivalent input circuit is given in Fig. 304. In this case, the input impedance h_i can certainly be neglected in comparison with the series resistor R_p, so that the latter is:

$$R_p = -V_{CC}/-I_B = 10/(12.8 \times 10^{-6}) = 800 \text{ k}\Omega.$$

From this we see that the series resistance increases with the load resistance.

Inductive load

Fig. 305 represents the transistor with an inductive load of negligible resistance. The operating conditions of the transistor are again chosen so that $-V_{CC} = 5$ V (point *A* in Fig. 306). In this case the collector-emitter voltage $-V_{CE}$ will be equal to the supply voltage $-V_{CC}$ in the absence of a signal.

The maximum output power of the transistor with a low value of Z_L
In this case the angle between the load line and the horizontal axis is given by $1/Z_L$. Here too, the transistor operating conditions must be chosen so that the voltage can move the same distance to both sides of the working point, without distortion occurring at maximum amplitude.

We shall assume that, without input signal, $-V_{CE} = -V_{CC} = 5$ V so that the constant collector current $-I_C = 5$ mA. The voltage can now move in both directions over a distance corresponding to 4.4 V, so that the peak-to-peak value of the collector-emitter voltage is 8.8 V. In this case, therefore, the maximum reactive power at the output is:

$$P_o = (\Delta V_{CE}/\sqrt{2}).(\Delta I_C/\sqrt{2}) = (4.4 \times 5/2)10^{-3} = 11 \times 10^{-3}\text{ W} = 11\text{ mW}.$$

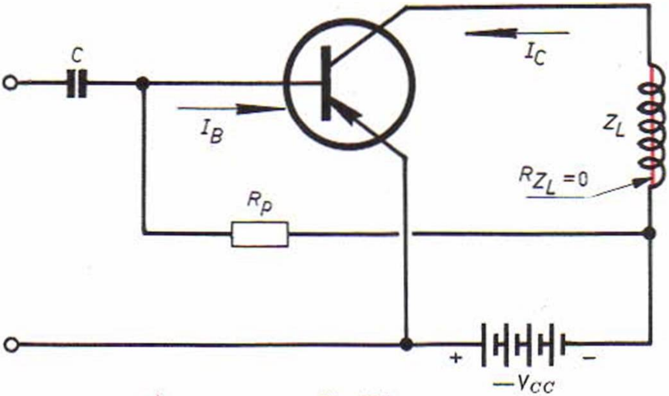

C
I_B
I_C
Z_L
$R_{Z_L}=0$
R_p
+
−
$-V_{CC}$

Fig. 305

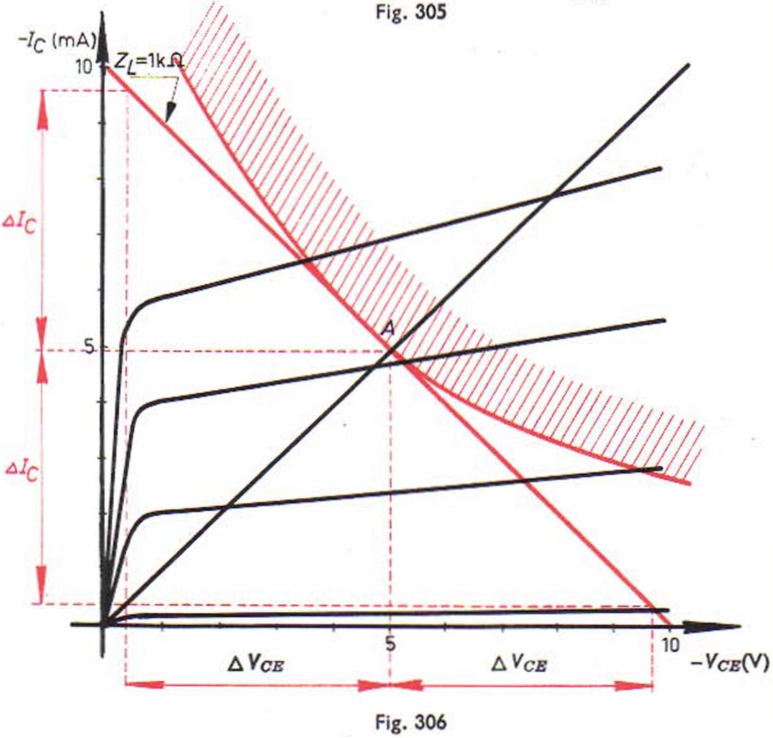


Fig. 306

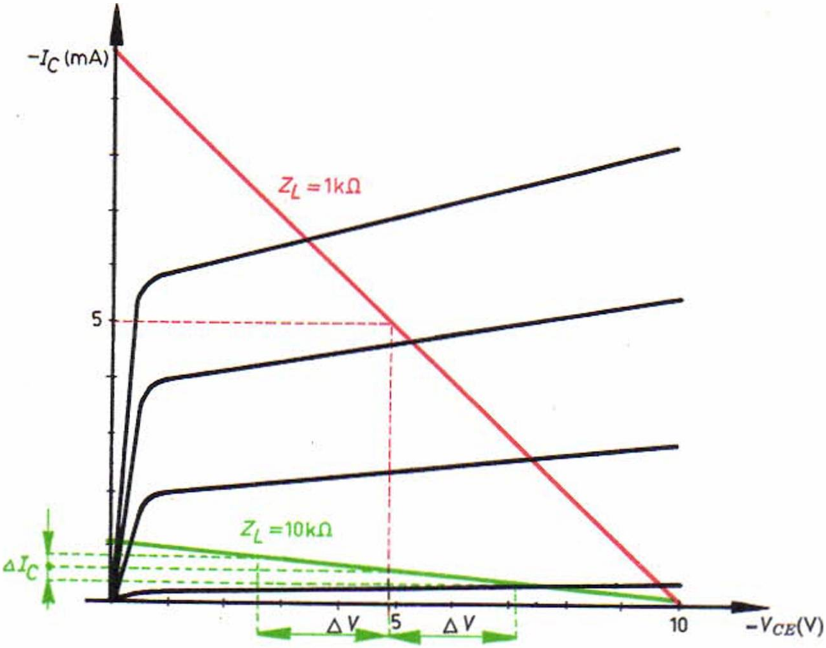
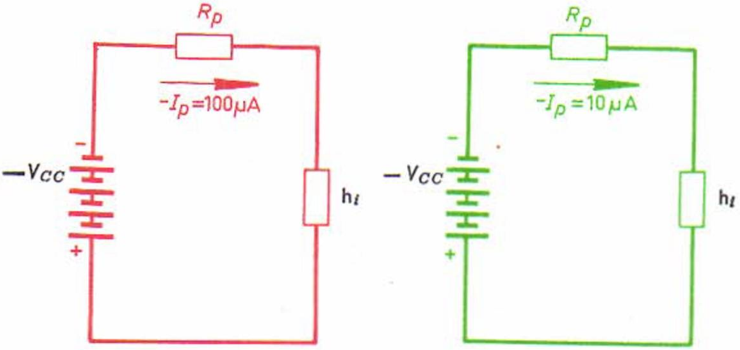


Fig. 307


R_p
$-I_p = 100\mu A$
$-V_{CC}$
-
+
h_i
R_p
$-I_p = 10\mu A$
$-V_{CC}$
-
+
h_i

(a) (b)

Fig. 308

Maximum output power of the transistor with a high value of Z_L

The slope of the load line decreases as Z_L increases. The green load line in Fig. 307 corresponds to a load impedance Z_L of 10 kΩ. In this case too the working point is chosen so that, in the absence of an input signal, $-V_{CE} = -V_{CC} = 5$ V. The collector-emitter voltage can again vary in both directions to a maximum value ΔV_{CE} of 4.4 V, corresponding to $I_C = 0.3$ mA.

The power available at the output is now:

$$P_o = (\Delta V_{CE}/\sqrt{2}) \cdot (\Delta I_C/\sqrt{2}) = (4.4 \times 0.3/2)10^{-3} = 0.66 \times 10^{-3} \text{ W} = 0.66 \text{ mW}.$$

The maximum output power thus increases as the load impedance decreases.

The effect of the load impedance on the transistor bias

At a low value of the load impedance e.g. $Z_L = 1$ kΩ, and neglecting $-I_{CEO}$ in relation to $-I_C$, the constant base current will equal:

$$-I_B = -I_C/h_{FE} = 5/50 = 0.1 \text{ mA} = 10^{-4} \text{ A}.$$

if the transistor operating conditions are again adjusted so that $-V_{CE} = -V_{CC} = 5$ V, corresponding to a constant collector current $-I_C$ of 5 mA, and the current amplification factor is 50. The resistance of the input circuit (Fig. 308a) can be calculated from the values of $-I_B$ and $-V_{CC}$:

$$R_p + R_i = V_{CC}/I_B = 5/10^{-4} = 5 \times 10^4 = 50 \text{ k}\Omega.$$

Since the input resistance R_i of the transistor may be neglected in comparison with the series resistance R_p, the latter may be taken as equal to 50 kΩ.

At a high value of the load impedance, e.g. $Z_L = 1$ kΩ, and the same transistor operating conditions, the constant base current will be:

$$-I_B = -I_C/h_{FE} = 0.5/50 = 0.01 \text{ mA} = 10^{-5} \text{ A}.$$

Fig. 308b represents the equivalent input circuit. Since, in this case also, the input resistance of the transistor may be neglected in relation to the series resistance, we have

$$R_p = V_{CC}/I_B = 5/10^{-5}\ \Omega = 500 \text{ k}\Omega.$$

28.3. Maximum power gain

In order to determine the maximum power gain we shall represent the transistor amplifier stage by the four-pole shown in Fig. 309. The power gain is now the quotient of the power available at the output and the power conveyed to the input. In the input circuit v_i represents the driving voltage and i_i the driving current. The power conveyed to the input is therefore:

$$P_i = v_i i_i.$$

In the output circuit v_o represents the available output voltage and i_o the current flowing through this circuit. The available output power is therefore:

$$P_o = v_o i_o.$$

The power gain is given by the equation:

$$G_P = P_o/P_i = v_o i_o/v_i i_i = (v_o/v_i) \, . \, (i_o/i_i).$$

In this equation v_o/v_i represents the voltage gain and i_o/i_i the current gain of the transistor:

$$G_V = v_o/v_i \text{ and } G_A = i_o/i_i \, .$$

This means that the power gain of a stage is equal to the product of the voltage gain and the current gain. In order to determine the effect of the output load on the power gain, we will now investigate what effect this load has on the voltage gain and then on the current gain. From this it is possible to determine at which value of the output load the maximum power gain can be obtained.

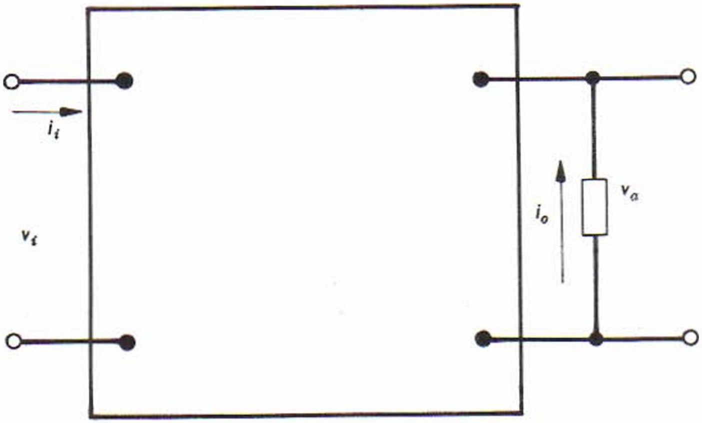


Fig. 309

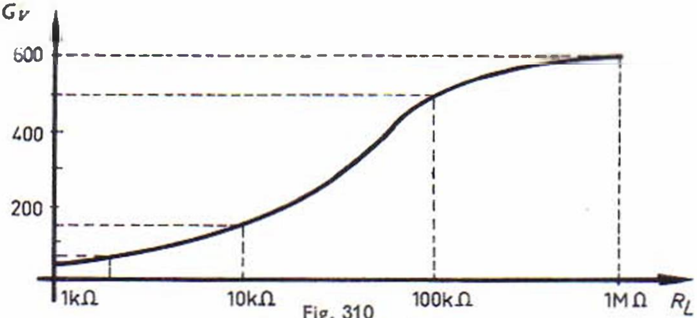


Fig. 310

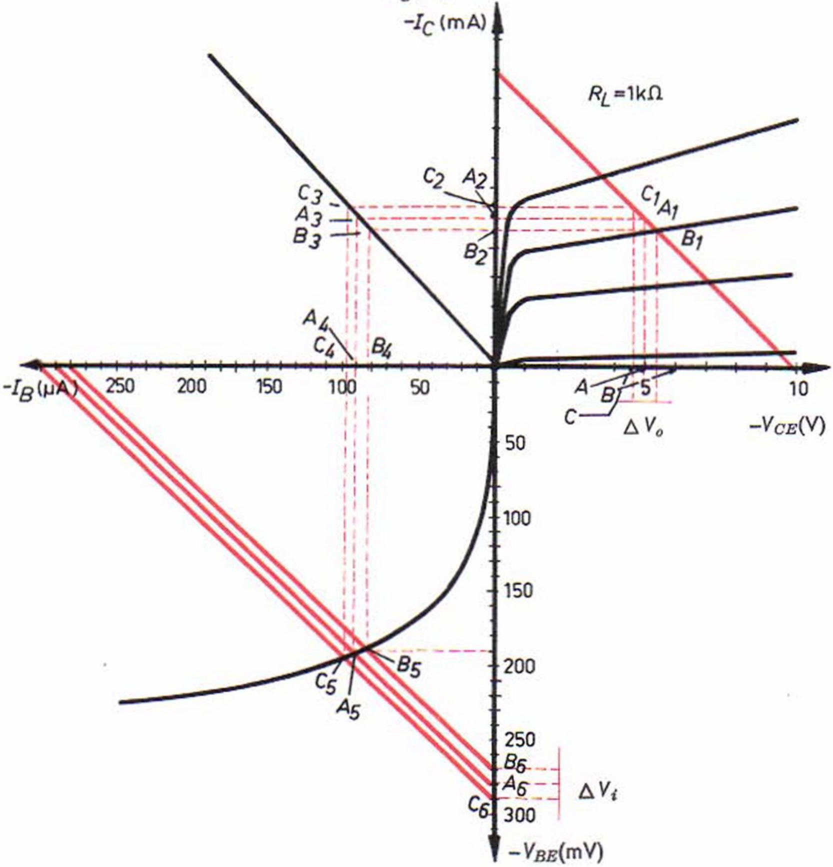


Fig. 311

28.4. The effect of the load impedance on the voltage gain

Fig. 310 shows the behaviour of the voltage gain as a function of the load impedance of the transistor. From this curve we see that the voltage gain $G_V = v_o/v_i$ is lowest at low values of the output impedance and then increases rapidly to reach a maximum value as this impedance approaches infinity. When $R_L = 0$ we have $G_V = 0$, as the output voltage v_o is equal to the product of $R_L = 0$ and the collector current.

Calculation of the voltage gain for a given load impedance

We will now calculate the voltage gain with the aid of the family of characteristics for a given load impedance. To this and we draw, in Fig. 311, the load line corresponding to $R_L = 1\ \text{k}\Omega$; the working point is chosen so that $-V_{CE} = -V_{CC}/2$ (point A_1 on the $-I_C = f(-V_{CE})$ characteristic, point A_3 on the $-I_C = f(-I_B)$ characteristic and point A_5 on the $-I_B = f(-V_{BE})$ characteristic). We will assume that the transistor is operating with matched drive. The peak-to-peak value of the input voltage $\hat{v}_i$ is represented by B_6C_6, while BC represents the peak-to-peak value of the output voltage $\hat{v}_o$:

$$G_V = \hat{v}_o/\hat{v}_i = BC/B_6C_6.$$

Under these conditions, the calculation of the voltage gain from the family of characteristics does not offer any particular difficulty.

28.5. The effect of the load impedance on the current gain

In Fig. 312 the current gain of the transistor is plotted as a function of the load impedance. This current gain is given by the quotient of the output current, i_o and the input current i_i, where i_o is assumed to be the peak-to-peak value of the collector current, and i_i the peak-to-peak value of the driving current.

The curve in Fig. 312 shows that the current gain is a maximum at low values of the load impedance, and gradually decreases with increasing value of this impedance; at an infinitely high load impedance, the current gain drops to zero.

To determine the current gain from the family of characteristics for a given load impedance, we set to work as follows. Fig. 313 represents the complete family of characteristics of the transistor. Let us assume that the load impedance, as in the previous case, is 1 kΩ. We will assume that $-V_{CE} = -V_{CC}/2$, which fixes the working point of the transistor (point A_1 in the $-I_C = f(-V_{CE})$ characteristic, point A_3 in the $-I_C = f(-I_B)$ characteristic and point A_5 in the $-I_B = \mathrm{f}(-V_{BE})$ characteristic). In this case the transistor is operating with matched drive. The peak-to-peak value of the driving current is given by B_6C_6 and that of the output current by B_2C_2, so that the current gain is equal to:

$$G_A = \hat{\imath}_o/\hat{\imath}_i = (B_2C_2)/(B_6C_6).$$

In this way, therefore, the current gain of this stage can easily be determined from the characteristics of the transistor if the load impedance and the generator impedance are known.

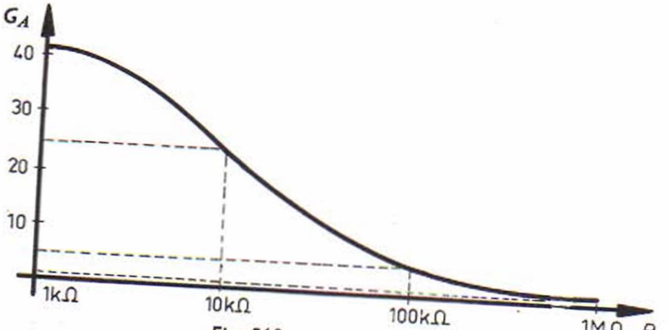


Fig. 312

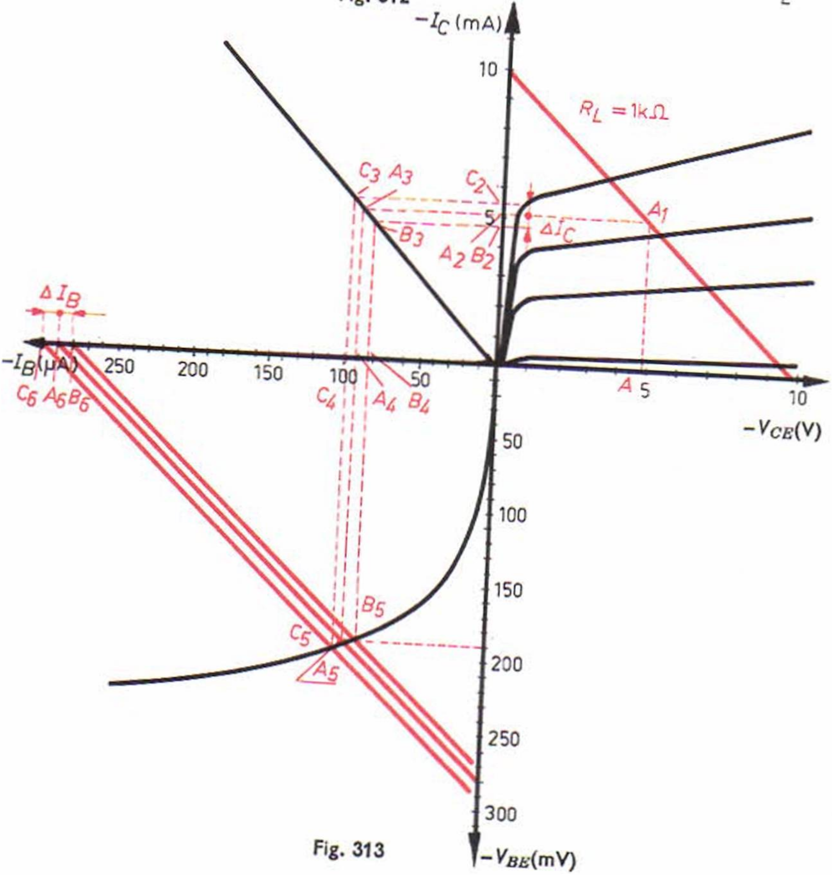


Fig. 313

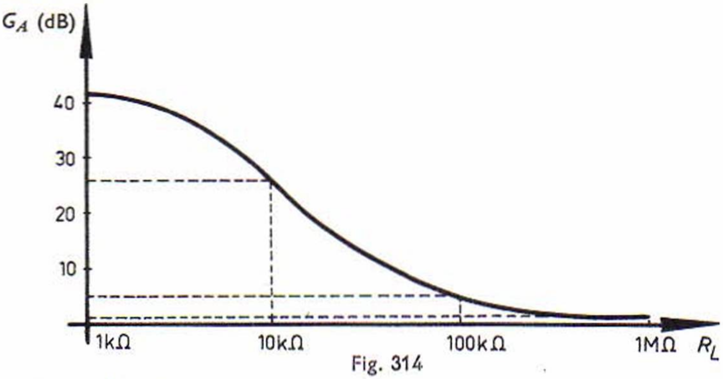


Fig. 314

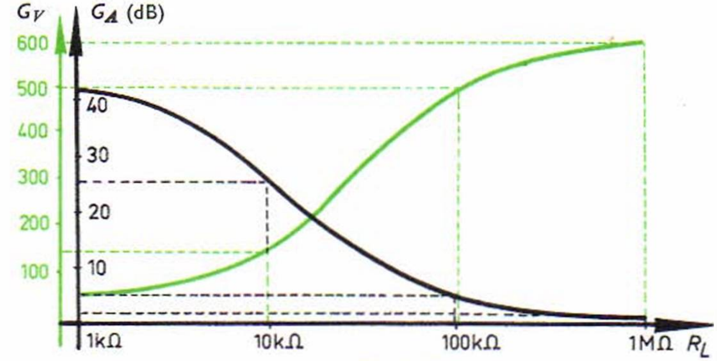


Fig. 315

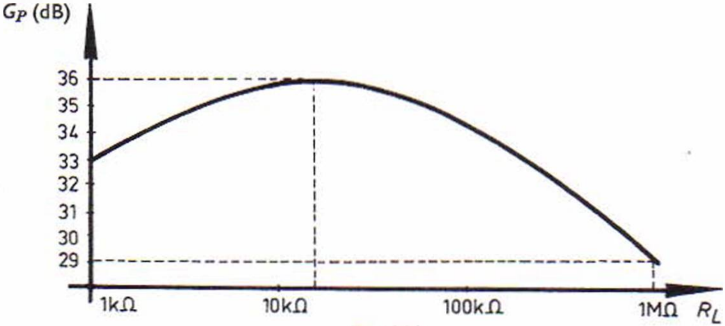


Fig. 316

28.6. The effect of the load impedance on the power gain

We will now represent both the voltage gain (printed green) and the current gain (printed black) graphically as functions of the load impedance (Figs. 314 and 315). As we explained on page 330, the power gain is equal to the product of the voltage gain and the current gain. Consequently, we can use two curves to determine the power gain as a function of the load impedance R_L (Fig. 316). To do this we plot the product of the voltage gain and the current gain for a number of values of the load impedance.

Load impedance of 1 kΩ

With a load impedance of 1 kΩ, the current gain is large and is almost equal to the maximum value, while the voltage gain is small. In this case, therefore, the power gain is:

$$G_P = G_V . G_A = 42 \times 50 = 2100 = 32.2 \text{ dB}.$$

With this value of the load impedance the power gain is therefore low.

Load impedance of 20 kΩ

With a load impedance of 20 kΩ the voltage gain is found to be 200 and the current gain 21, so that the power gain is:

$$G_P = G_V . G_A = 200 \times 21 = 4200 = 36 \text{ dB}.$$

The power gain is the maximum at this value of R_L. This must be ascribed to the fact that the output impedance of the transistor in grounded emitter is also of the order of 20 kΩ.

Load impedance of 1 MΩ

The voltage gain here approximates to its maximum value of 600, but the current gain is only slightly greater than 0. The power gain is given by:

$$G_P = G_V . G_A = 600 \times 1.5 = 900 = 29.5 \text{ dB}.$$

It follows from this that the power gain for a load impedance of 1 MΩ is only small.

Fig. 317 shows that the power gain G_P is a maximum if the load impedance is equal to the output impedance of the transistor (20 kΩ). The power gain drops rapidly with increasing or decreasing values of the load impedance. Consequently there is an important difference between the maximum power gain and the maximum power which a transistor can supply.

In Fig. 318 the red load line corresponds to the maximum power gain; the slope of this line tan $A = 1/R_L$ corresponds to a value of R_L equal to the output impedance h_o of the transistor. By contrast, the black load line corresponds to the maximum power which is available at the output of the transistor; the slope of this line is determined by the constant output-power hyperbola.

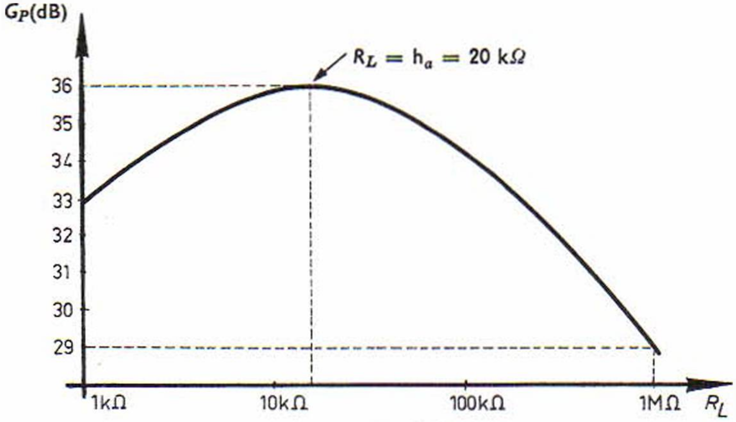


Fig. 317

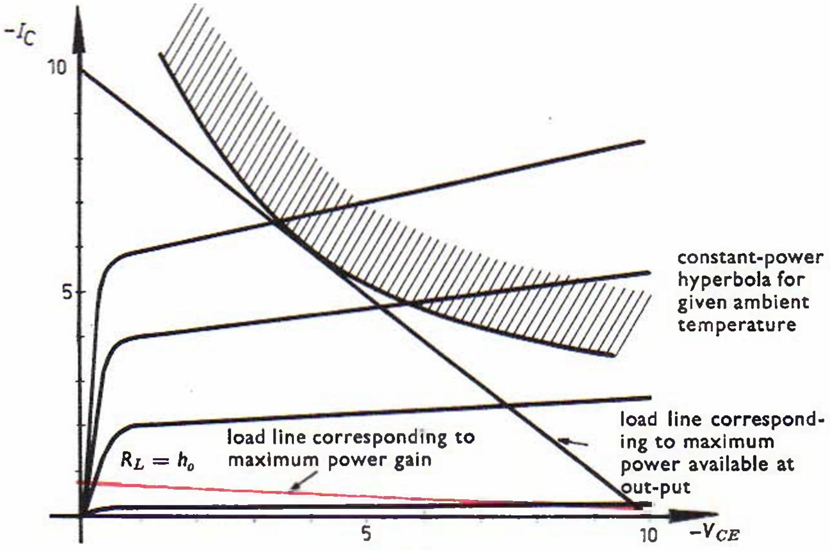


Fig. 318

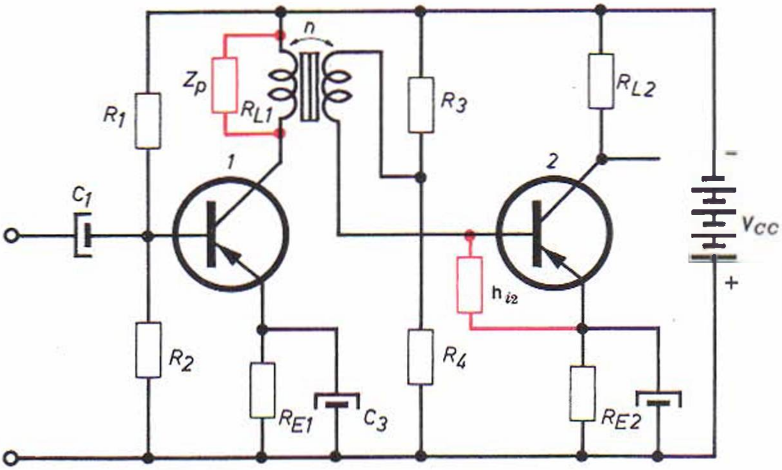

n
Zp
RL1
R1
R2
C1
1
RE1
C3
R3
R4
hi2
2
RL2
RE2
–
+
VCC

Fig. 319

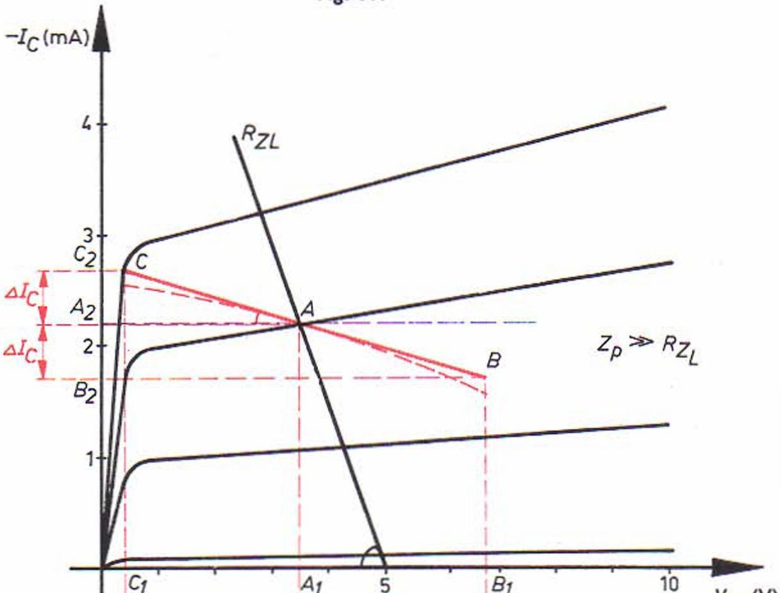


Fig. 320

Transistors connected in cascade

Transistors connected in cascade can be coupled either by means of a transformer or by an *RC* circuit.

29.1. Transformer-coupled stages

Fig. 319 represents a circuit in which two transistors are coupled by means of a transformer. The $-I_C = \mathrm{f}(-V_{CE})$ characteristic of transistor 1 is shown in Fig. 320. Assuming that the supply voltage $-V_{CC}$ is equal to 5 V, the d.c. load line of the transistor can be drawn through this point. Since the primary of the transistor has a negligible resistance this line will have a very steep slope. In drawing this load line, however, the resistance in series with the emitter must be taken into account. The position of the working point on this load line can now be determined in the absence of an input signal.

For alternating current the load impedance of transistor 1 is much greater than zero, as it equals the input impedance of transistor 2 transferred to the primary. If the transformation ratio is n and the input impedance of transistor 2 is equal to h_{i2}, then this transferred impedance is:

$$Z_p = n^2 \mathrm{h}_{i2}\,.$$

The load impedance for alternating current is therefore always much greater than that for direct current, so that the load line for alternating current (printed red in the figure) will always have a much flatter slope than the load line for direct current. The maximum power which transistor 1 can supply to transistor 2 depends on one hand on the voltage across the output terminals of transistor 1 and on the other hand on the current flowing through the circuit. (Projections of the voltage swing on the load line and on the $-V_{CE}$ - and $-I_C$ axes:

so that $v_{o\ \mathrm{peak}} = A_1B_1$; $\qquad i_{o\ \mathrm{peak}} = A_2B_2$,

so that $P_{o1} = (v_{o\ \mathrm{peak}}/\sqrt{2}).\ (i_{o\ \mathrm{peak}}/\sqrt{2}) = (A_1B_1) \cdot (A_2B_2)/2.$

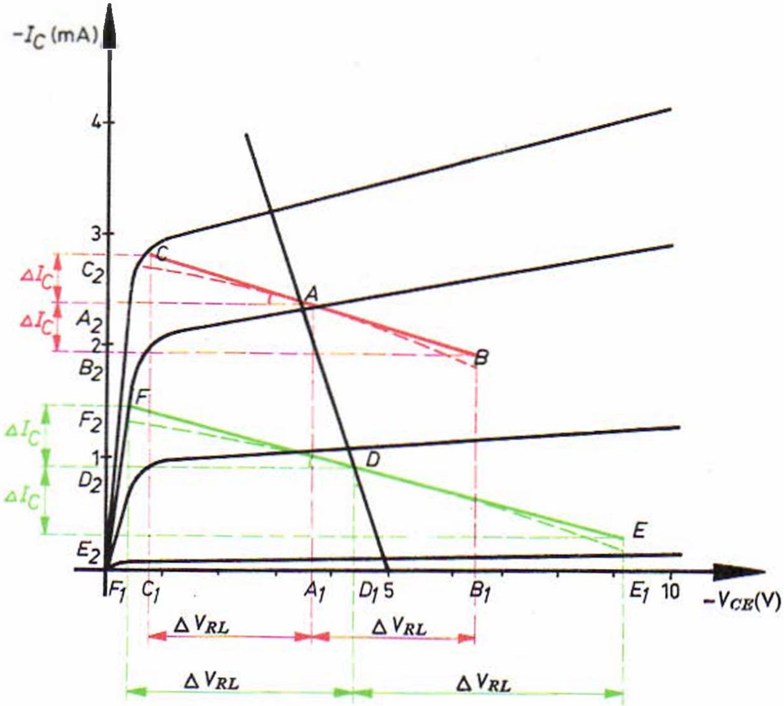
In order to obtain the maximum undistorted power we must therefore ensure that A_1B_1 is equal to A_1C_1.

We will now move the working point on the d.c. load line to point D (see Fig. 321). The angle between the green a.c. load line and the horizontal axis remains unchanged. The maximum power which transistor 1 can supply to transistor 2 is now:

$P_{o1} = (v_{o\ \mathrm{peak}}/\sqrt{2}).\ (i_{o\ \mathrm{peak}}/\sqrt{2}) = (D_1E_1)\ .\ (D_2E_2)/2.$

This power is greater than the power which we have just calculated while, in addition, the constant collector current is lower. The efficiency of this amplifier is therefore considerably greater than in the previous case.

In these considerations, we have assumed that the input impedance of transistor 2 is constant, whatever the voltage variations. This is indeed true for small signals, but the variation in the input impedance cannot be neglected for large signals. This is because this impedance increases when the base becomes less negative (point F of the region DF), and decreases if the base becomes more negative (point E of the region DE). The dotted line represents the actual a.c. load line.



$-I_C$ (mA)
4
3
2
1
ΔI_C
ΔI_C
ΔI_C
ΔI_C
C_2
A_2
B_2
F_2
D_2
E_2
C
A
B
F
D
E
F_1
C_1
A_1
D_1
5
B_1
E_1
10
$-V_{CE}$(V)
ΔV_{RL}
ΔV_{RL}
ΔV_{RL}
ΔV_{RL}

Fig. 321

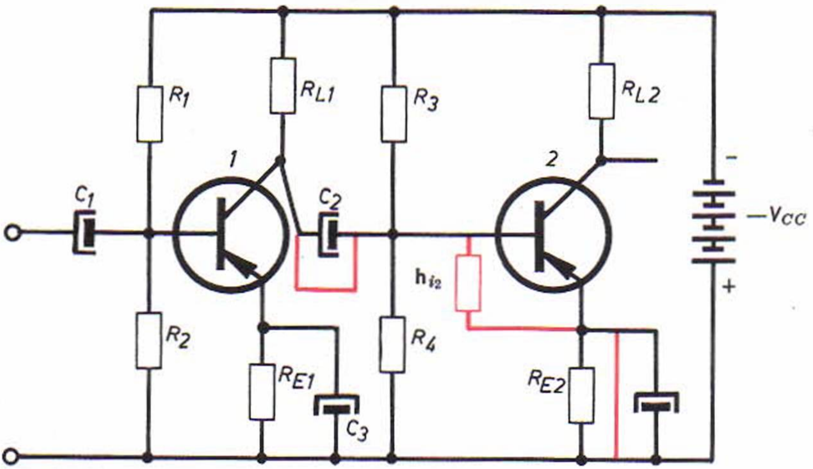

R1
RL1
R3
RL2
1
2
C1
C2
hi2
R2
RE1
C3
R4
RE2
−
−VCC
+

Fig. 322

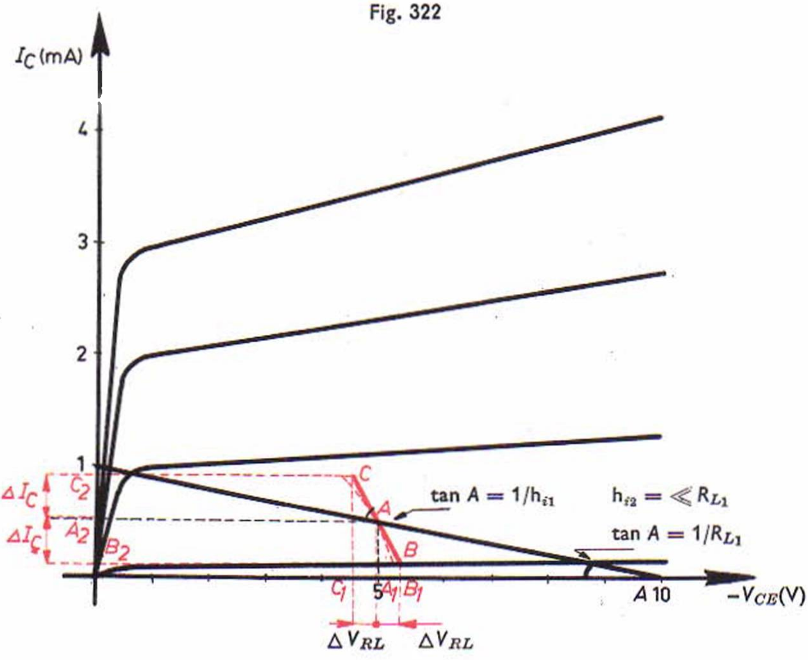


Fig. 323

29.2. *RC*-coupled stages

Fig. 322 represents a two-stage amplifier in which transistors 1 and 2 are coupled by an *RC* circuit. The load of transistor 2 is also a resistance. We again draw the d.c. load line in the $-I_C = f(-V_{CE})$ characteristic of transistor 1 (Fig. 323). This is determined by the supply voltage $-V_{CC} = 10$ V and the load resistance $R_{L1} = 10\ \mathrm{k}\Omega$ of transistor 1. As in the previous case, we select the working point so that $-V_{CE} = -V_{CC}/2$.

For alternating current, capacitor C_2 presents a negligible impedance, so that we may assume the input impedance of transistor 2 to be connected directly in parallel with the load resistance of transistor 1. Now this input impedance is only a few hundreds of ohms, i.e. much lower than the load impedance. This means that the load impedance of transistor 1 is much lower for alternating current than it is for direct current.

In Fig. 323 the a.c. load line of transistor 1 is printed red. The power available at the output is:

$$P_{o1} = (\Delta V_o/\sqrt{2})\,(\Delta i_o/\sqrt{2}).$$

The peak value of the output voltage in the direction of decreasing current is here equal to A_1B_1, and since the chosen working point must be situated at the centre of the voltage swing, C_1A_1 will be equal to A_1B_1.

From the projections of point C on the $-I_C$ axis and on the $-V_{CE}$ axis (see Fig. 324) we have:

$$P_{o1} = (v_{o\ peak}/\sqrt{2}).\,(i_{o\ peak}/\sqrt{2}) = (A_1B_1)\,.\,(A_2B_2)/2.$$

In this case the output voltage is only a subordinate factor, the principle object being to convey the maximum possible power to the transistor. In the first place this power depends on the peak value A_2B_2 of the collector current.

If we select the working point on the d.c. load line so as to give a greater constant collector current and a smaller collector-emitter voltage, e.g. point D (printed green in the figure), the peak value D_2E_2 of the collector current will increase considerably. As the load lines at points A and D have the same slope, the voltage amplitude will scarcely decrease as a result of this shift of the working point:

$$i_{c\ peak} = D_2E_2; \qquad v_{ce\ peak} = D_1E_1$$

so that

$$P_{o1} = (v_{o\ \mathrm{peak}}/\sqrt{2}).\,(i_{o\ \mathrm{peak}}/\sqrt{2}) = (D_1E_1)\,.\,(D_2E_2)/2.$$

We see that in this case transistor 1 can supply a greater power to transistor 2. In practice the operating conditions will be chosen such that the collector-emitter voltage in the absence of a signal is of the order of 1 to 2 V. The load resistance of transistor 1 will be made as large as possible, so that most of the collector alternating current will flow through the input impedance of transistor 2. In these discussions it was assumed, in setting out the a.c. load lines, that the input impedance of transistor 2 is constant, which is in fact true for small signals. For large signals the voltage swing in the directions of points F and E on the load line, resulting from the input signal, will mean that the input impedance of transistor 2 increases if the voltage swings in the direction of F, and decreases if the voltage swings in the direction of E. The actual load line is shown dotted in the figure.

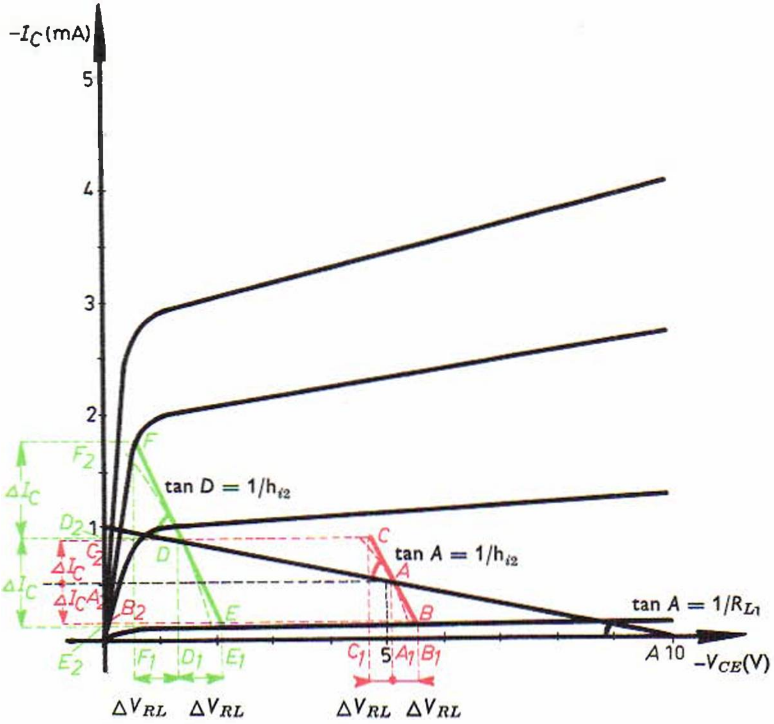
$-I_C$ (mA)
5
4
3
2
1
F
F_2
ΔI_C
tan $D = 1/h_{i2}$
D_2
D
C_2
ΔI_C
ΔI_C
ΔI_C
A_2
B_2
E
E_2
F_1
D_1
E_1
C
tan $A = 1/h_{i2}$
A
B
tan $A = 1/R_{L1}$
C_1
5
A_1
B_1
A 10
$-V_{CE}$(V)
ΔV_{RL}
ΔV_{RL}
ΔV_{RL}
ΔV_{RL}

Fig. 324

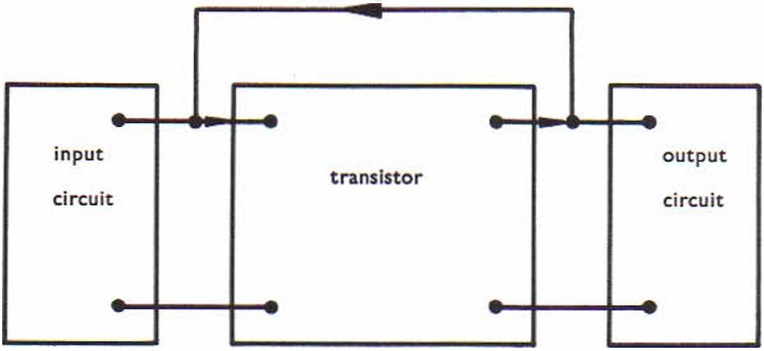

input
circuit
transistor
output
circuit

Fig. 325

CHAPTER 30

The use of transistors at radio frequencies

The behaviour of transistors in the amplification of r.f. signals is undoubtedly much more complicated than it is in the amplification of a.f. signals. Some parameters which were neglected in previous chapters will prove to be of great importance; these relate principally to the internal capacitances of the transistor.

The upper limit to the frequency at which a transistor can be used is determined by the transit time of the electrons. There is in fact a limit to the velocity with which these can move in a semiconductor, so that there is sometimes a limit to the amplification of a transistor at very high frequencies. The signals to be amplified at radio frequencies are usually weak and the principal object is to obtain the greatest possible power gain. To achieve this, the impedance of the signal source must equal the input impedance of the transistor, while the load impedance must equal its output impedance: in this case the transistor is operating with matched drive. However, this is not possible for all r.f. transistors, because matching may give rise to instability. In this case, a certain amount of "mismatching" must be applied purposely, with the resulting loss of amplification. The relatively smooth shape of the curve in Fig. 317 illustrates this.

To simplify the following explanations, we will first limit ourselves to regarding the transistor as an active element, in other words, we shall examine the factors which determine the amplification at radio frequencies, and shall investigate the variations to which this amplification is subject. We shall then regard the transistor as a passive element in order to discover the influence of the transistor input on the driving circuit, and of the output on the load circuit, and we shall also discuss the negative feedback from the output circuit to the input circuit (see Fig. 325).

In order to obtain an appreciation of the difficulties which can be encountered in the use of transistors in a wide frequency range, it is necessary, for the sake of completeness, to discuss the variations of the various parameters. As the greatest power gain is achieved in the common emitter configuration, we shall limit ourselves to this.

CHAPTER 31

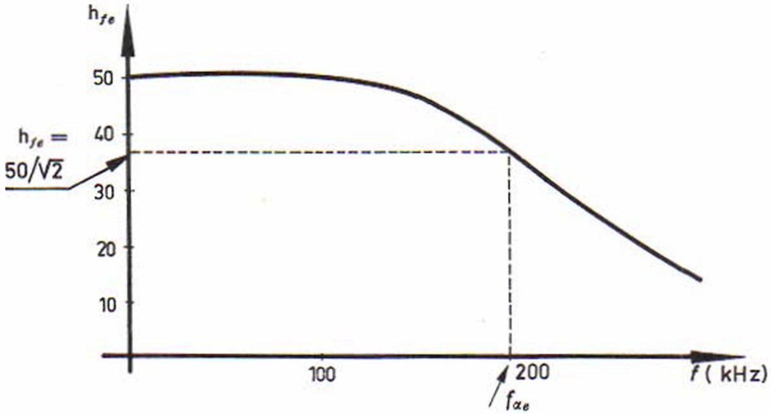
The transistor as an active element

R.F. signals are preferably amplified with matched drive, so that the transistor is both current – and voltage – driven. The principal factor in current drive is the current amplification factor of the transistor, represented by the symbol h_{fe}, but for voltage drive, with which we are principally concerned in connection with r.f. applications, the slope y_{fe} of the transistor is the most important factor (also see page 198). Unfortunately, both these parameters vary with the frequency of the signal to be amplified.

The frequency at which the current gain has dropped to $1/\sqrt{2}$ of its original value at low frequencies (see Fig. 326), is termed the cut-off frequency $f_{\alpha e}$ of the transistor. By analogy, it would be possible to determine the cut-off frequency f_{ye} at which the slope has dropped to $1/\sqrt{2}$ of its original low frequency value (Fig. 327). In practice, however, $f_{\alpha e}$ is usually much lower than f_{ye}, so that the latter does not have to be taken into account. For this reason the manufacturer only publishes the cut-off frequency $f_{\alpha e}$, so that f_{ye} will have to be determined by measurement. The r.f. amplification of a transistor stage is determined by the following four important parameters:

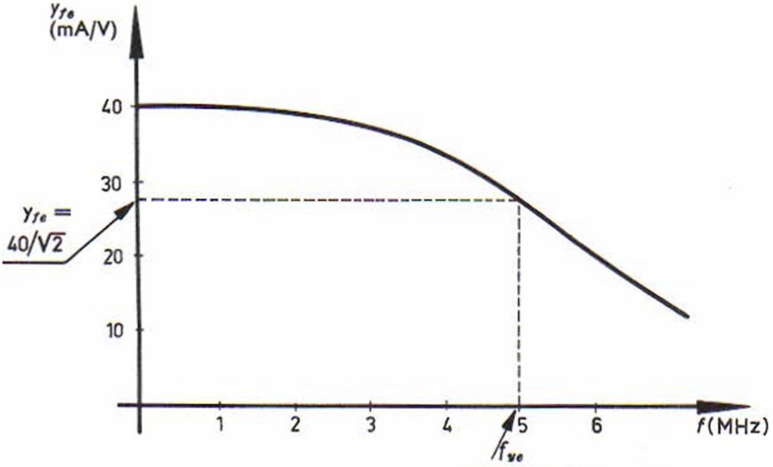
The current amplification factor.
The cut-off frequency of the current-amplification factor.
The slope.
The cut-off frequency of the slope.

We will now discuss the relationship between these quantities and the operating conditions of the transistor.



Current-gain cut-off frequency

Fig. 326



Slope cut-off frequency

Fig. 327

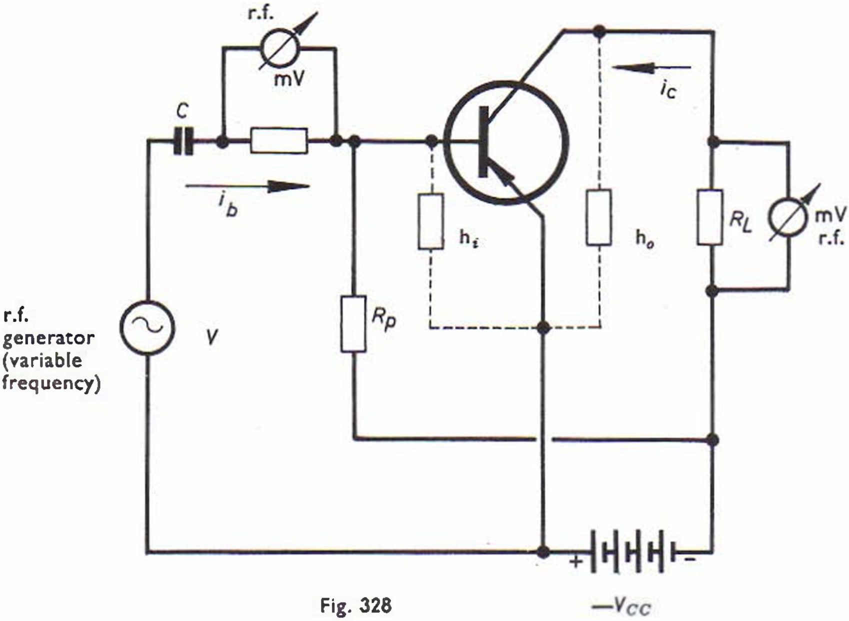

r.f.
mV
C
i_b
i_c
h_i
h_o
R_L
mV
r.f.
r.f. generator (variable frequency)
V
R_p
+
−
$-V_{CC}$

Fig. 328

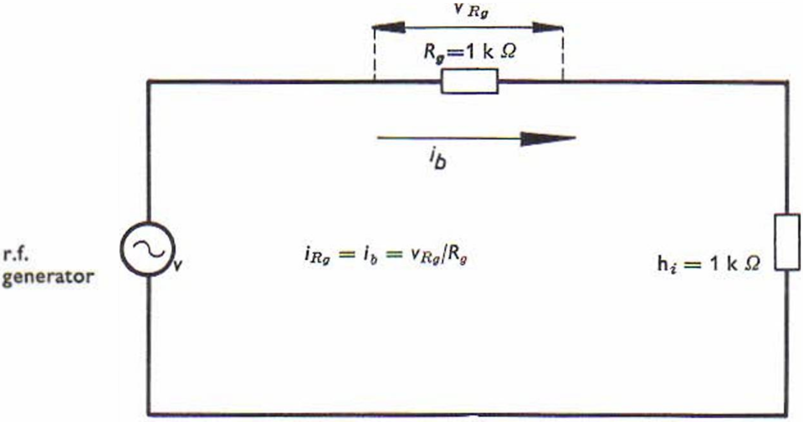

v_{R_g}
$R_g = 1\ \text{k}\,\Omega$
i_b
r.f. generator
v
$i_{R_g} = i_b = v_{R_g}/R_g$
$h_i = 1\ \text{k}\,\Omega$

Fig. 329

31.1. Current-amplification factor

We have already discussed the influence of the collector current and of the collector-emitter voltage on the current-amplification factor on page 193. For a low-power transistor this influence is only small.

We will now examine the effect which the frequency of the input signal has on the current-amplification factor. To do this we start from the circuit of Fig. 328. The transistor is operating with matched drive, that is, the generator impedance R_g is equal to the impedance h_i of the transistor. In addition the transistor is loaded by a resistor R_L equal to the output impedance h_o of the transistor, for example, $R_L = h_o = 20\ \mathrm{k}\Omega$. We shall assume that the resistance of the r.f. generator itself is zero and that there is a resistor R_g connected in series with it, of value equal to the input impedance of the transistor. R.F. valve voltmeters are connected across the terminals of the resistor R_g and over those of resistor R_L.

Fig. 329 represents the equivalent circuit of the input.
This consists of an r.f. generator loaded by the generator impedance R_g in series with the input impedance h_i of the transistor. Suppose that $R_g = h_i = 1\ \mathrm{k}\Omega$. The voltage across resistor R_g can be measured by means of the valve voltmeter across it. From the values of this resistance and this voltage follows the current in the input circuit, i.e. the base current of the transistor:

$$i_{R_g} = i_b = v_{R_g}/R_g.$$

The equivalent output circuit is given in Fig. 330. The value of the collector current can be determined from the voltage measured across the load impedance:

$$i_c = v_{RL}/R_L.$$

Once the collector current i_c and the base current i_b have been determined in this way they give the current gain of the transistor:

$$G_A = i_c/i_b.$$

By varying the frequency of the generator these measurements can be carried out for various frequencies, thus making it possible to plot the curve representing the current gain as a function of the frequency (see Fig. 331). This curve shows that the current gain is practically constant for low frequencies, and decreases fairly rapidly at higher frequencies.

31.2. Cut-off frequency

In Fig. 331 the value of the current gain which determines the cut-off frequency is particularly important. This is given by the point A on the curve of current gain plotted against frequency; this point corresponds to a current gain equal to $1/\sqrt{2}$ of the original value at low frequencies.

We will now investigate the effect of the collector current on the cut-off frequency determined in this way. To do this we carry out the same measurements for various values of the collector current and plot the frequencies at which this current is $1/\sqrt{2}$ lower than the original value, as a function of the collector current. Fig. 332 represents the curve obtained in this way. This shows that the cut-off frequency remains practically constant for values of the collector current between 0.25 mA and 3 mA, which means that the current gain of the transistor always drops to $1/\sqrt{2}$ of its original value at a certain frequency.

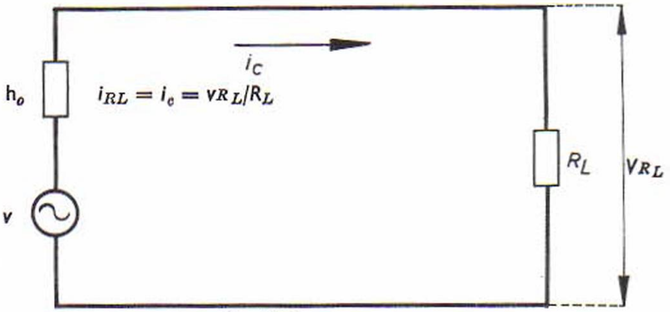

h_o
i_C
$i_{RL} = i_c = v_{RL}/R_L$
v
R_L
v_{RL}

Fig. 330

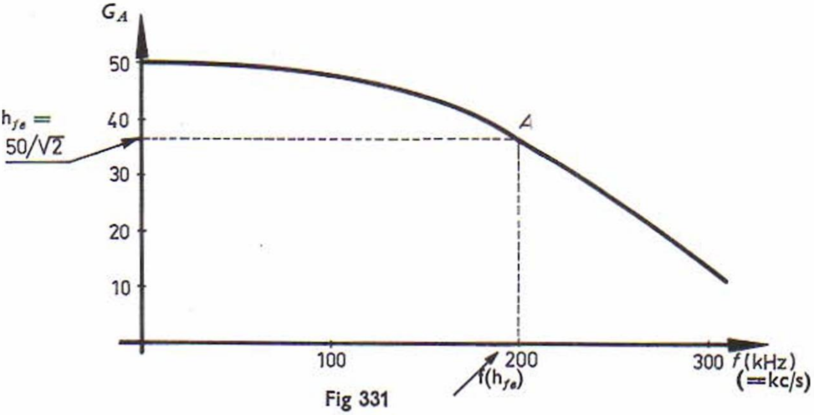


Fig 331

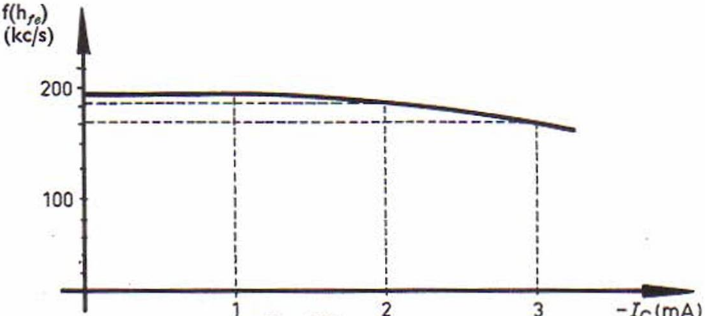


Fig. 332

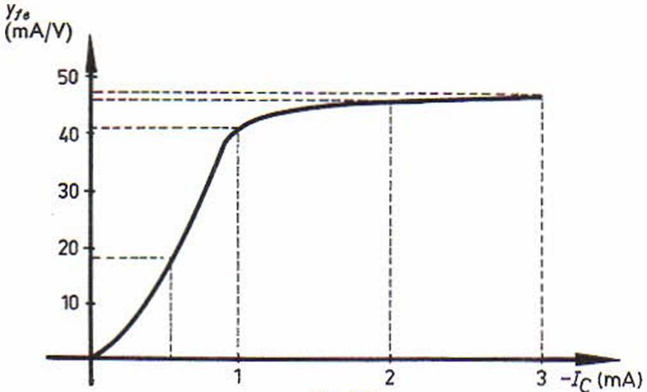


Fig. 333

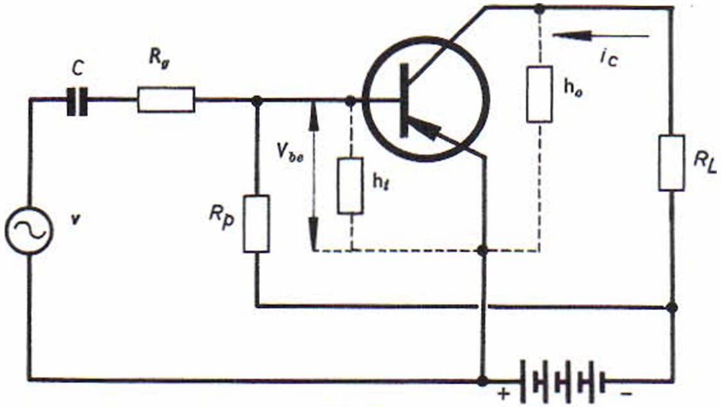

C
R_g
v
R_p
V_{be}
h_i
h_o
i_C
R_L
+
−
$-V_{CC}$

Fig. 334

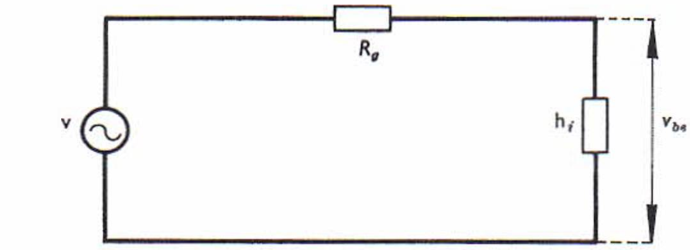


Fig. 335

31.3. Slope

The influence of the collector current on the slope of the transistor has already been discussed on page 202. The slope was found to increase with the collector current (see Fig. 333).

We shall now investigate to what extent the slope of the transistor is affected by the frequency of the input signal. Let us assume that in this case also the transistor is operated with matched drive, so that the generator impedance is equal to its input impedance, while the load impedance equals the output impedance of the transistor (Fig. 334). We will again take the resistance of the generator itself as zero, but with a separate resistor $R_g = h_i = 1\ \text{k}\Omega$ connected in series with it. Fig. 335 represents the equivalent input circuit. This contains an r.f. generator connected in series with the generator impedance R_g and the input impedance h_i of the transistor. We can again determine the voltage across the input impedance of the transistor, i.e. the base-emitter voltage v_{be} by means of a valve voltmeter.

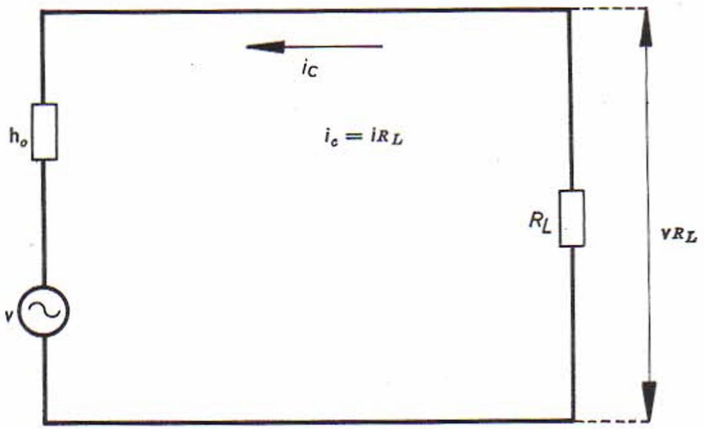
The slope is given by the quotient of the variations of the collector current and the variations of the base-emitter voltage causing these current variations.

$$y_{fe} = \Delta i_c / \Delta v_{be}.$$

Fig. 336 represents the equivalent output circuit of the transistor. The voltage across the load impedance R_L can be measured with the aid of an r.f. valve voltmeter. The current through this circuit, i.e. the collector current i_c, can be calculated from the value of this impedance:

$$i_{R_L} = i_c = V_{R_L}/R_L.$$

When the slope of the transistor has been determined in this way the measurement can be repeated for other frequencies, so that we obtain a curve representing the slope as a function of the frequency of the input signal (see Fig. 337). Point A on this curve, corresponding to the frequency at which the slope of the transistor has decreased by a factor $1/\sqrt{2}$ relative to its original value at low frequencies is particularly important. This is the cut-off frequency f_{ye} for the slope of the transistor. This cut-off frequency will be found to be appreciably higher than the cut-off frequency for the current gain, which explains why a transistor can operate very well at higher frequencies than the cut-off frequency $f_{\alpha e}$ which is published by the manufacturer (for the common-emitter configuration).



i_C
$i_c = i_{R_L}$
h_o
R_L
v_{R_L}
v

Fig. 336

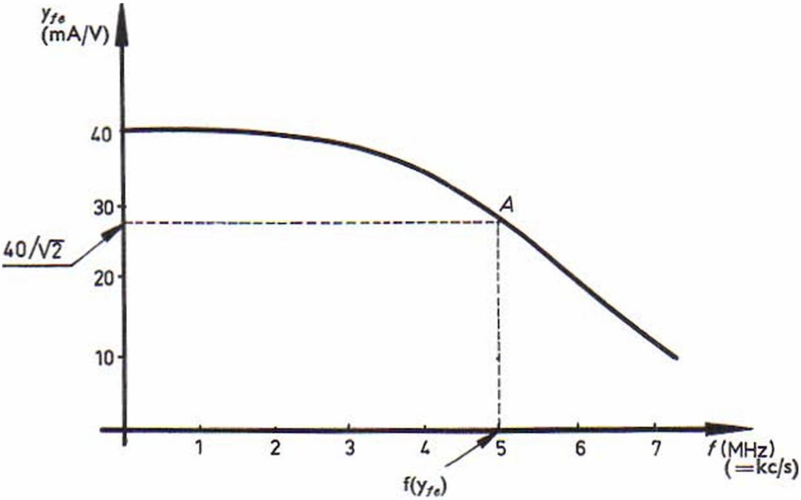


Fig. 337

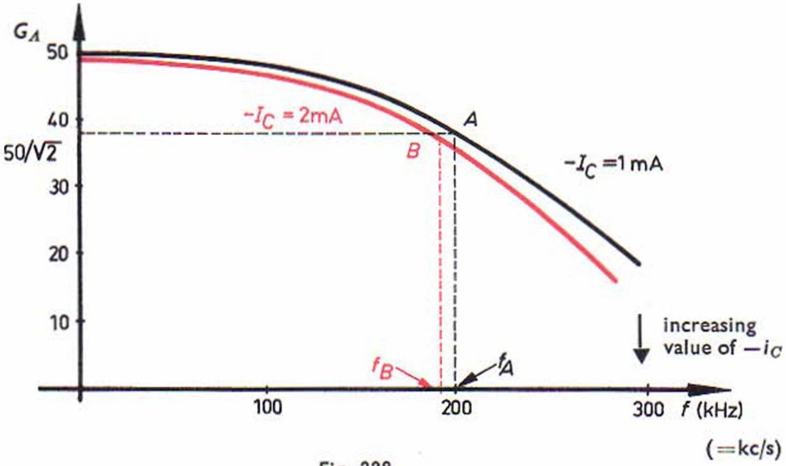

G_A
50
40
$50/\sqrt{2}$
30
20
10
$-I_C = 2\text{mA}$
A
B
$-I_C = 1\,\text{mA}$
increasing value of $-i_C$
f_B
f_A
100
200
300
f (kHz)
(=kc/s)

Fig. 338

31.4. Determination of the optimum value of the constant collector current

We will now attempt to find the optimum value of the constant collector current, in order to ensure satisfactory operation of the transistor at high frequencies. Assuming that the transistor operates with matched drive, it is desirable that both the current gain and the slope should be as large as possible at the frequencies at which the transistor is used, or in other words that both cut-off frequencies should be as high as possible.

The cut-off frequency for the current-amplification factor

In order to determine the cut-off frequency for the current-amplification factor we will plot the current gain at a constant collector current of 1 mA as a function of the frequency (Fig. 338). The cut-off frequency is now given by the point A on this curve, whose projection on the ordinate corresponds to $1/\sqrt{2}$ of the maximum current gain, and which gives the point f_A on the abscissa as the cut-off frequency. If we now increase the collector-current, to $-I_C = 2$ mA for example, the same value of the cut-off frequency is obtained at a point f_B which differs very little from f_A. As far as current drive is concerned, therefore, the operation of a transistor at high frequencies is found to be practically independent of the operating conditions.

The cut-off frequency for the slope

We will again assume that the constant collector current is $-I_C = 1$ mA. The slope of the transistor is plotted as a function of the frequency in Fig. 339. The projection of the point A on the abscissa gives the point f_A as the cut-off frequency and its projection on the ordinate gives a point corresponding to $1/\sqrt{2}$ of the slope at low frequencies. If we allow the constant collector current to increase, the same value of the slope will be obtained at a frequency f_B which is lower than f_A. With voltage drive the transistor works less satisfactorily as the collector current is increased, so if it is required to use the transistor at the highest possible frequency it is recommended that this current should be made as small as possible. In practice there is a lower limit to the value of this current, represented by the leakage current, and principally by the variations of the leakage current as a function of the temperature.

In the $-I_C = f(-V_{CE})$ characteristic of Fig. 340 the constant collector current $-I_C = 0.5$ mA. An increase in temperature will cause this current to increase, either because the $-I_c = f(-V_{ce})$ characteristic shifts, or because the transistor operating conditions move in the direction indicated.

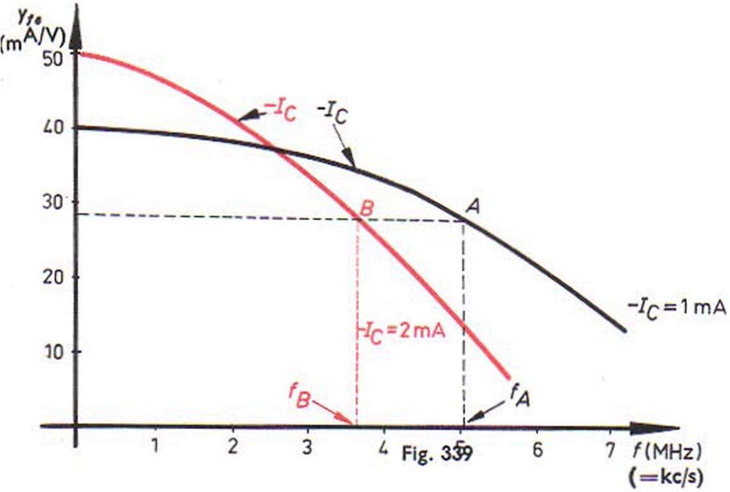


Fig. 339

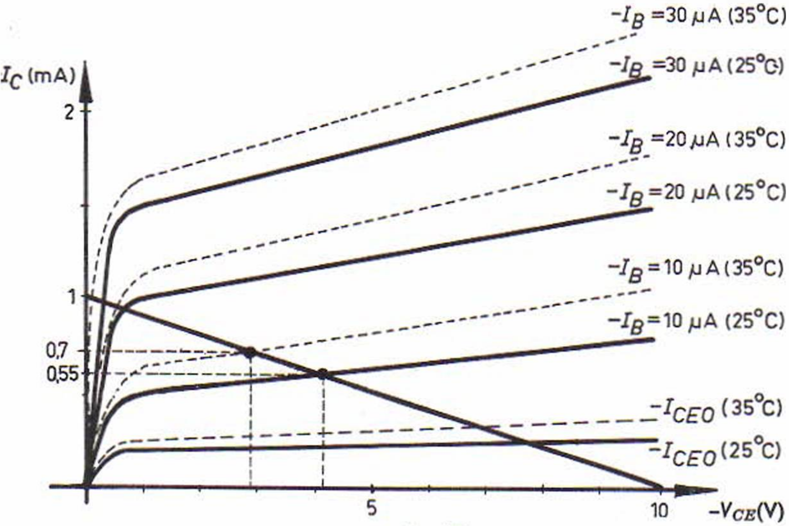


Fig. 340

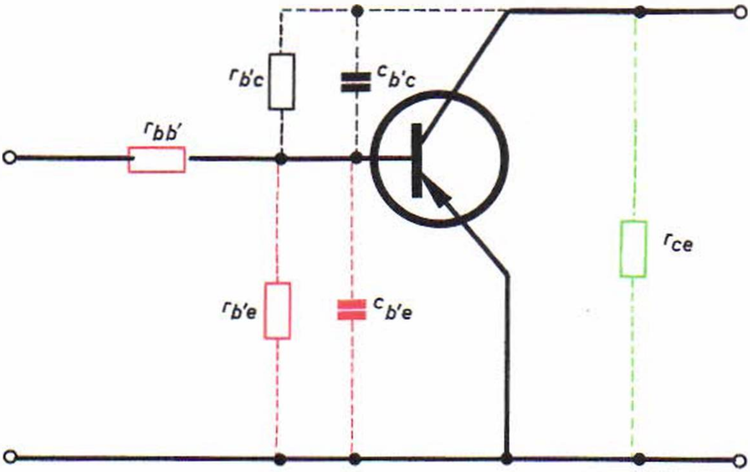

$r_{b'c}$
$c_{b'c}$
$r_{bb'}$
r_{ce}
$r_{b'e}$
$c_{b'e}$

Fig. 341

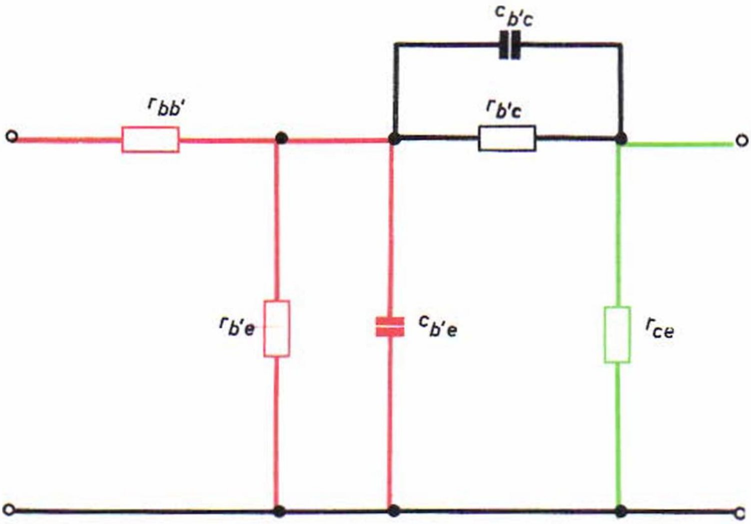

$c_{b'c}$
$r_{bb'}$
$r_{b'c}$
$r_{b'e}$
$c_{b'e}$
r_{ce}

Fig. 342

CHAPTER 32

The transistor as a passive element

The object of this chapter is to investigate the influence of the input impedance of the transistor on the circuit driving it, and of the output impedance on the load circuit. We shall also discuss in more detail the feedback which the transistor introduces between the load circuit and the input circuit. To this end we must define the various internal elements (resistances and capacitances) of the transistor more closely and must investigate to what extent these depend on the collector current and on the frequency.

32.1. Equivalent circuit

The input circuit is formed by the base-emitter circuit and the output circuit by the collector-emitter circuit, while the feedback is determined by the collector-base circuit. Fig. 341 shows the various transistor parameters which have more than a negligible influence on its behaviour at high frequencies.

Fig. 342 represents the equivalent circuit of the transistor. The elements printed in red represent the input circuit of the transistor, consisting of a resistance $r_{bb'}$, which must be ascribed to the region between the base connection and the base region of the junctions, a resistance $r_{b'e}$ in series with r_{bb}, and which must be ascribed to the base-emitter junction, and a capacitance $c_{b'e}$, which must be ascribed to the diffusion of the holes in the base (the input capacitance of the transistor) and which is connected in parallel with $r_{b'e}$.

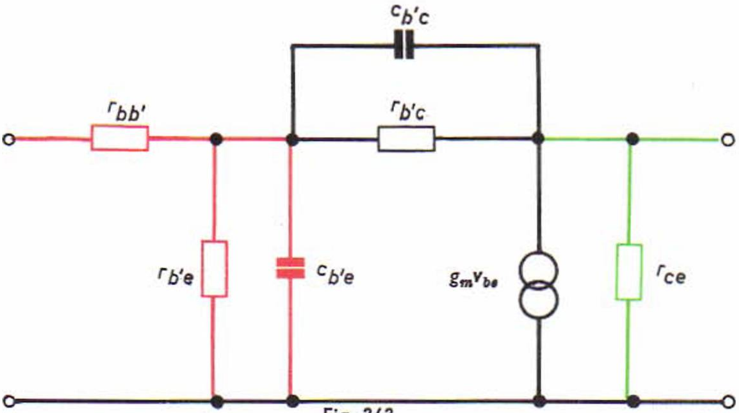
The output circuit (printed green in the figure) contains only a resistance r_{ce} which represents the output impedance h_o of the transistor.

The feedback circuit (printed black in Fig. 343) comprises a resistance $r_{b'c}$ which must be ascribed to the collector-base junction and a capacitance $c_{b'c}$ which must also be ascribed to this junction and which is connected in parallel with the resistance $r_{b'c}$. The equivalent circuit obtained in this way refers only to the passive elements of the transistor, so that we must add the active elements in the form of a current generator $g_m v_{b'e}$.

In the usual transistor terminology, the resistances $r_{b'e}$, $r_{b'c}$ and r_{ce} are designated by the symbols $g_{b'e}$, $g_{b'c}$, g_{ce} respectively, corresponding to the reciprocal resistances, i.e.:

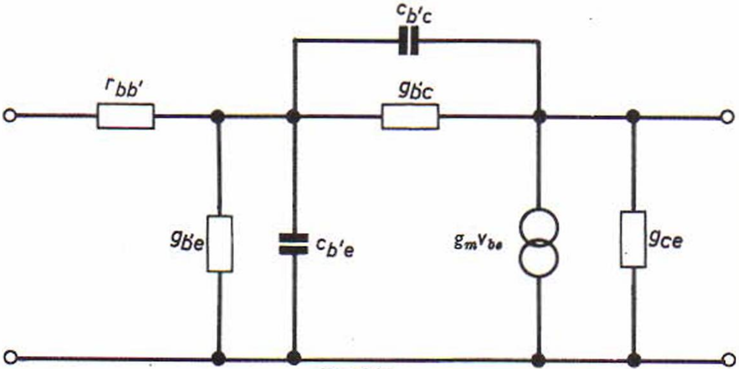
$g_{b'e} = 1/r_{b'e}$ represents the input admittance of the transistor.
$g_{b'c} = 1/r_{b'c}$ represents the admittance of the internal feedback.
$g_{ce} = 1/r_{ce}$ represents the output admittance of the transistor.

The equivalent circuit, as represented in Fig. 344, is the same as Fig. 342, the only difference being that the symbols have been adapted to this terminology. The equivalent circuit can also be represented more simply, as in Fig. 345. The input circuit consists of resistances and capacitances, so that these can be replaced by r_{ieq} parallel with a capacitance c_{ieq}. This is also true for the output circuit, which can be represented by a resistance r_{oeq} in parallel with a capacitance c_{oeq}. Finally the feedback circuit is formed by a capacitance c_{req} in series with a resistance r_{req}.



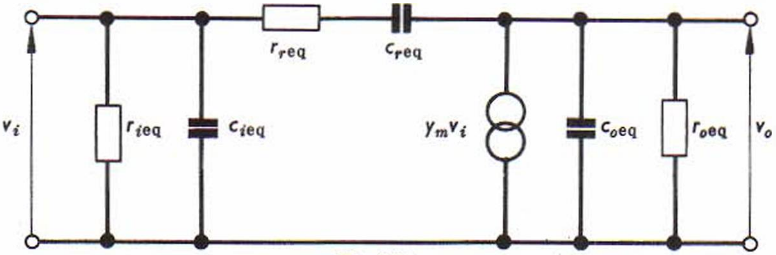
$c_{b'c}$
$r_{bb'}$
$r_{b'c}$
$r_{b'e}$
$c_{b'e}$
$g_m v_{be}$
r_{ce}

Fig. 343



$c_{b'c}$
$r_{bb'}$
$g_{b'c}$
$g_{b'e}$
$c_{b'e}$
$g_m v_{be}$
g_{ce}

Fig. 344



r_{req}
c_{req}
v_i
r_{ieq}
c_{ieq}
$y_m v_i$
c_{oeq}
r_{oeq}
v_o

Fig. 345

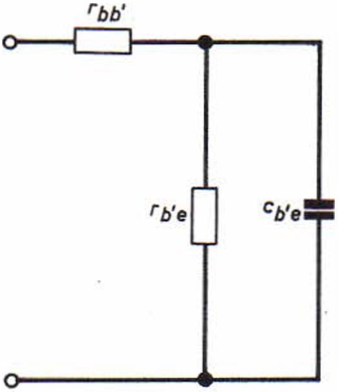


Fig. 346

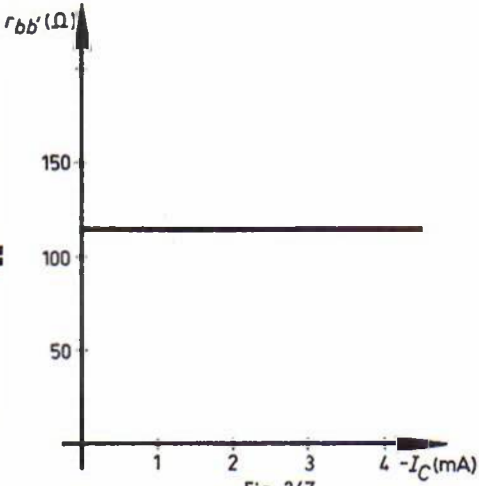


Fig. 347

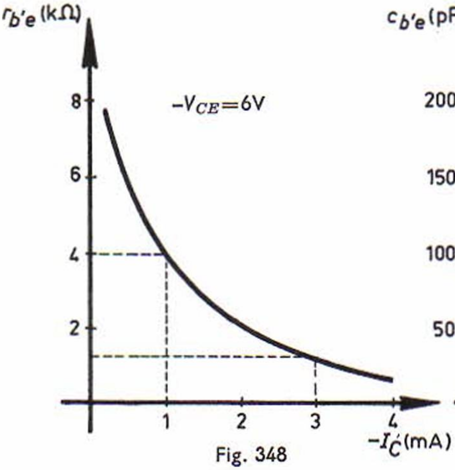


Fig. 348

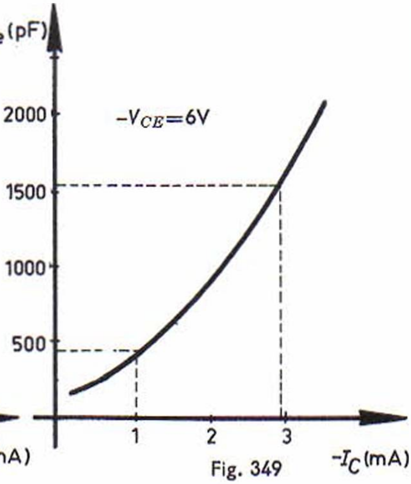


Fig. 349

CHAPTER 33

The influence of the collector current and of the frequency on the various r.f. parameters of the transistor

33.1. The input circuit

The input circuit consists of the resistance r_{bb}' in series with the resistance $r_b{}'_e$ in parallel with which is connected the capacitance $c_b{}'_e$ (see Fig. 346).

The influence of the collector current on r_{bb}'

In Fig. 347 r_{bb}' is plotted as a function of the collector current $-I_C$. The figure shows that this resistance does not vary with the collector current. This is explained by the fact that the collector current only flows through the collector-base and base-emitter junctions and thus has no effect on the resistance which must be ascribed to the base connections.

The influence of the collector current on $r_{b'e}$

The resistance $r_b{}'_e$ corresponds to the input impedance of the transistor; the relationship between this impedance and the collector current has already been discussed on page 186 and is represented graphically in Fig. 348. This figure shows that $r_b{}'_e$ decreases steeply as the collector current increases.

The influence of the collector current on $c_{b'e}$

The curve in Fig. 349 shows the input capacitance $c_b{}'_e$ of the transistor plotted as a function of the collector current. This capacitance is found to increase sharply with the value of the collector current. We must point out that this capacitance is relatively high, i.e. of the order of several hundred picofarads; this point must therefore be carefully considered in the design of a circuit for amplifying r.f. signals.

With the aid of the curves that we have just described we can find out how the equivalent resistance r_{ieq} and capacitance c_{ieq} which represent the resistive and capacitive effects of the transistor input, depend on the collector current. For this purpose we will make use of the simplified representation of the transistor input which is given in Fig. 350. This consists of a resistor r_{ieq} connected in parallel with a capacitor c_{ieq}. With the aid of this representation we can determine the changes in value of these two elements as functions of the collector current.

The influence of the collector current on r_{ieq}

In Fig. 351 r_{ieq} is plotted as a function of the collector current. The variation of this resistance with the collector current shows great agreement with that of the base-emitter resistance $r_{b'e}$. We see that r_{ieq} also decreases rapidly with increasing collector current.

With the aid of this curve we can accurately determine the value of the input resistance of the transistor, i.e. the resistance which is connected in parallel with the driving circuit, for each value of the collector current. As the calculation of this resistance falls outside the scope of this book, we will content ourselves with stating its value at the end of this chapter as a function of:

$$g_{bb'} = 1/r_{bb'},\ g_{b'e} = 1/r_{b'e},\ c_{b'e} \text{ and } c_{b'c}.$$

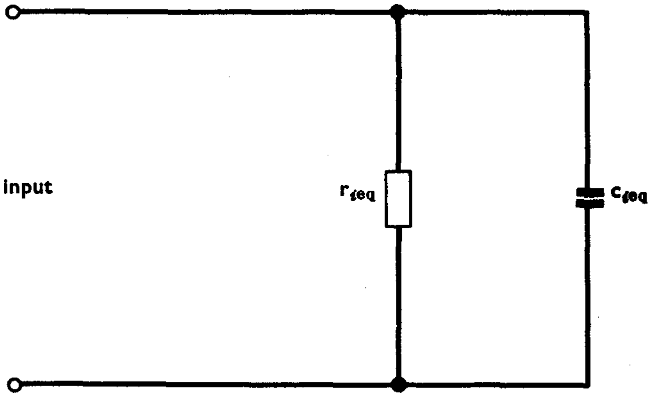

input
r_{ieq}
c_{ieq}

Fig. 350

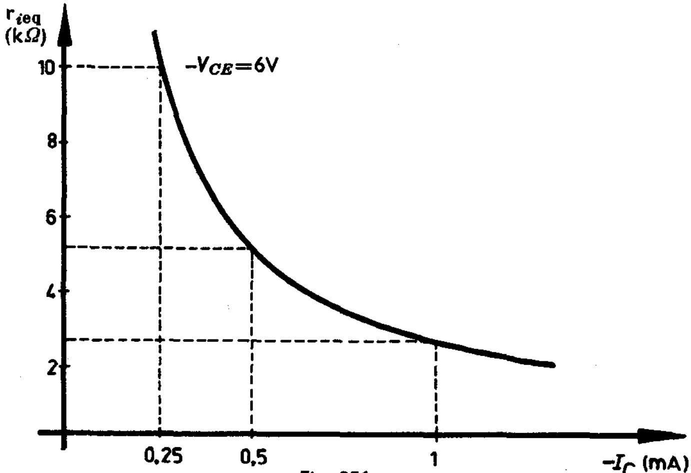


Fig. 351

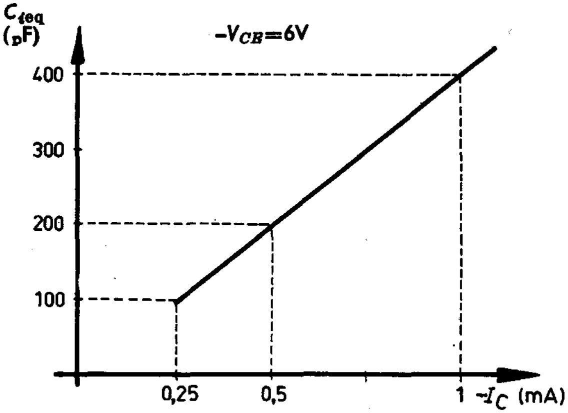


Fig. 352

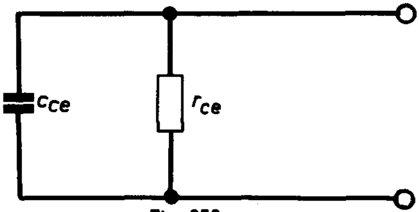


Fig. 353

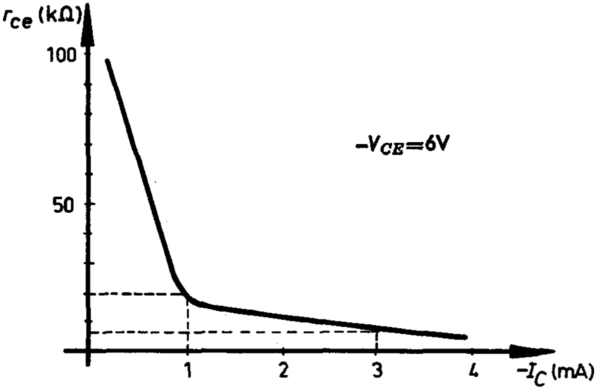


Fig. 354

The influence of the collector current on c_{ieq}

The capacitance c_{ieq} is principally determined by the input capacitance of the transistor, and its variation as a function of the collector current shows great agreement with that of $c_{b'e}$ (see Fig. 352). We see from this that c_{ieq} increases approximately linearly with the collector current.

We can determine this capacitance accurately for a given value of the collector current, and in this way we can determine the capacitance which is connected across the input circuit. The value of c_{ieq} as a function of $c_{b'c}$, $c_{b'e}$ and $g_{bb'}$ is also given at the end of this chapter.

33.2. The output circuit

The output circuit consists of the collector-emitter resistance r_{ce} connected in parallel with a capacitance c_{ce} (see Fig. 353). We will now investigate the influence of the collector current and the collector-emitter voltage on these two parameters.

The influence of the collector current on r_{ce}

The collector-emitter resistance r_{ce} corresponds to the output impedance of the transistor, which has already been discussed in detail on page 189. Fig. 354 shows this resistance plotted as a function of the collector current; the resistance decreases rapidly with increasing collector current. At a constant collector current of 0.25 mA, it is still approximately 150 kΩ, but at a constant collector current of 3 mA the resistance of an OC 44 transistor has already dropped to 40 kΩ.

The influence of the collector-emitter voltage on r_{ce}

The influence of the collector-emitter voltage on the output impedance $h_o = r_{ce}$ of the transistor has already been discussed on page 189. It is found that this resistance increases only slightly with the collector-emitter voltage.

The influence of the collector current on c_{ce}

From the curve of Fig. 356, in which the output capacitance c_{ce} is plotted as a function of the collector current, we see that this capacitance increases rapidly with the current. This effect can easily be explained since the capacitance in question consists of the base-collector capacitance $c_{b'c}$, and the base-emitter capacitance $c_{b'e}$ connected in series. An increase in the collector current causes the base-collector junction to decrease in width with the result that the capacitance of this junction increases with the collector current. The output capacitance of the transistor is given by the equation:

$$1/c_{ce} = 1/c_{b'c} + 1/c_{b'e},$$

and increases with the value of $-I_C$.

The influence of the collector-emitter voltage on c_{ce}

The output capacitance c_{ce} is plotted in Fig. 357 as a function of the collector-emitter voltage. We see that this capacitance decreases as the collector-emitter current increases, but that the influence of the collector current is only slight.

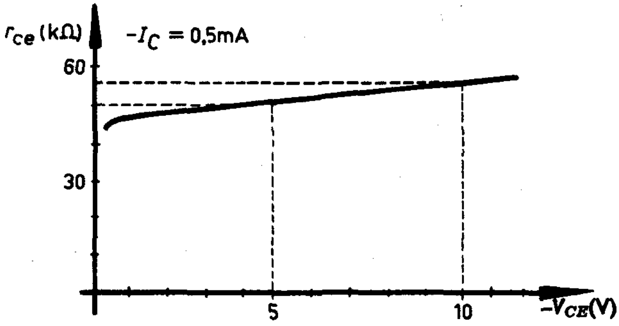


Fig. 355

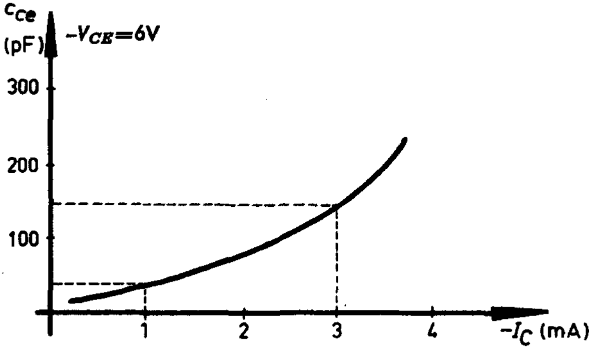


Fig. 356

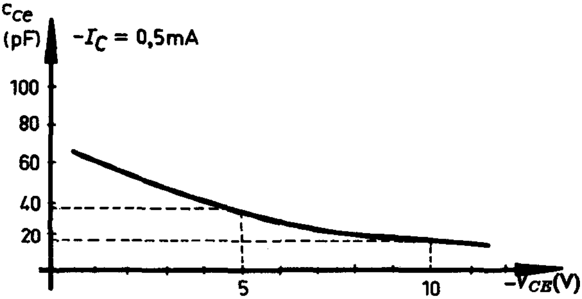
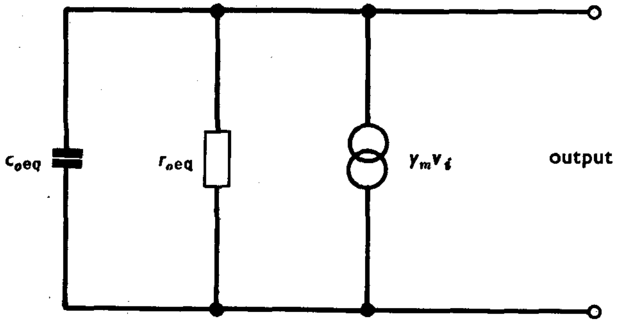


Fig. 357



c_{o}eq
r_{o}eq
$y_m v_i$
output

Fig. 358

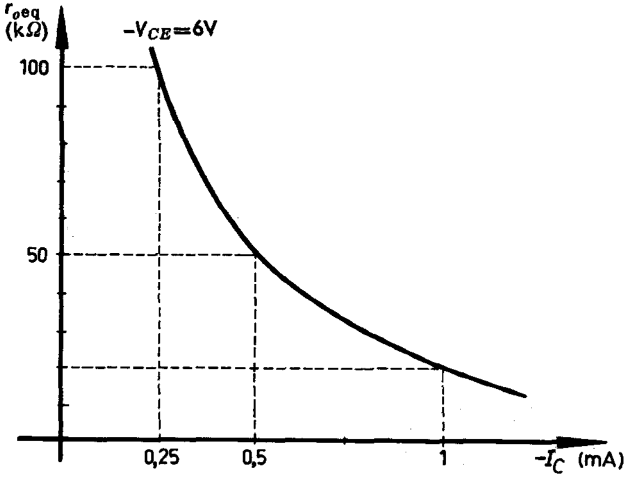


Fig. 359

The output citcuit can be represented by the equivalent circuit of Fig.358 in which an equivalent resistance $r_{o\mathrm{eq}}$ is connected in parallel with an equivalent capacitance $c_{o\mathrm{eq}}$. The former represents the purely resistive effect to be ascribed to the output of the transistor while the latter represents the purely capacitive effect of the output. Starting from the curves which we have just discussed, we can now determine the influence of the collector current and of the collector-emitter voltage on these two elements.

The influence of the collector current on $r_{o\mathrm{eq}}$

Fig. 359 shows the equivalent output resistance $r_{o\mathrm{eq}}$ of a transistor at high frequencies as a function of $-I_C$. From this curve it is possible to make an accurate determination of the value of this resistance for a given collector current, in order that it may be taken into account in the design of a circuit. For the sake of completeness the relationship between $r_{o\mathrm{eq}}$ and the transistor parameters:

$g_{ce} = 1/r_{ce}$, $g_{b'c} = 1/r_{b'c}$, $g_{bb'} = 1/r_{bb'}$, $g_{b'e} = 1/r_{b'e}$, $c_{b'c}$ and $c_{b'e}$.

is given at the end of this chapter.

In place of the impedance r_{ce} its reciprocal, the conductance $g_{ce} = 1/r_{ce}$ is often used. As we see from Fig. 360, the output impedance r_{ce} is connected in parallel with the load impedance. The equivalent value of these two impedances is given by:

$$1/r_{o\mathrm{eq}} = 1/r_{ce} + 1/R_L.$$

For the calculation it is easier to work with the conductances, that is with $g_{o\mathrm{eq}}=1/r_{o\mathrm{eq}}$, $g_{ce} = 1/r_{ce}$ and $g_L = 1/R_L$, so that we can simply write:

$$g_{o\mathrm{eq}} = g_{ce} + g_L.$$

The influence of the collector-emitter voltage on $r_{o\mathrm{eq}}$

The relationship between the collector-emitter voltage and the equivalent resistance $r_{o\mathrm{eq}}$ is represented by the curve in Fig. 361. The shape of this curve is practically identical with that which represents the collector-emitter resistance as a function of $-V_{CE}$; here too, the equivalent resistance increases slightly with the collector-emitter voltage.

The influence of the collector current on $c_{o\mathrm{eq}}$

The variation of the equivalent capacitance $c_{o\mathrm{eq}}$ as a function of the collector current $-I_C$ is represented by the curve of Fig. 362. We see that this capacitance increases with the collector current; the explanation for this has already been given. At the end of this chapter this capacitance is expressed in terms of the various transistor parameters already referred to.

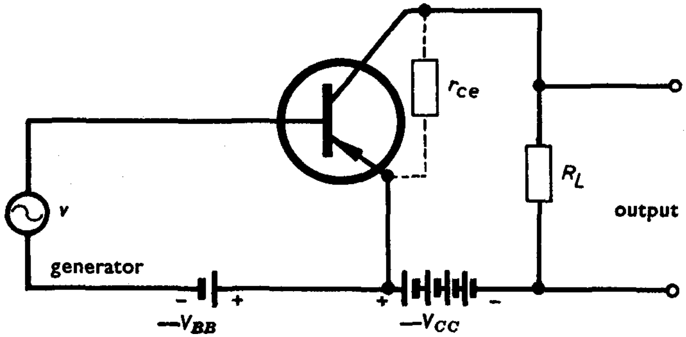

r_{ce}
R_L
v
output
generator
−
+
+
−
$—V_{BB}$
$—V_{CC}$

Fig. 360

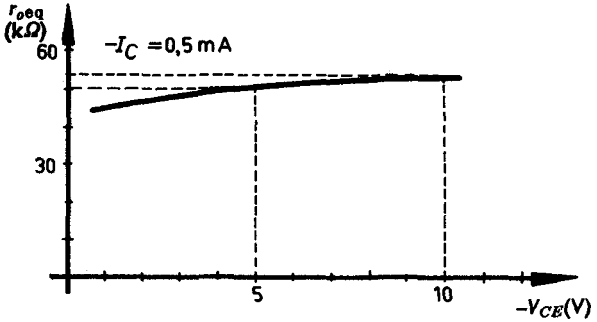


Fig. 361

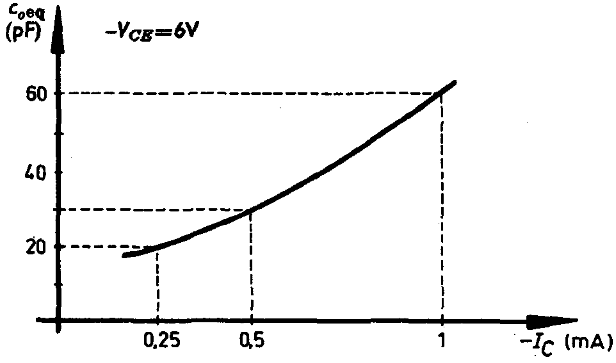


Fig. 362

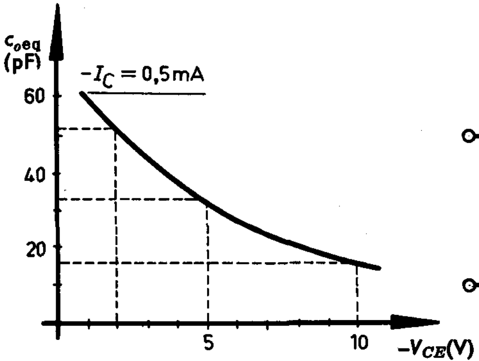


Fig. 363

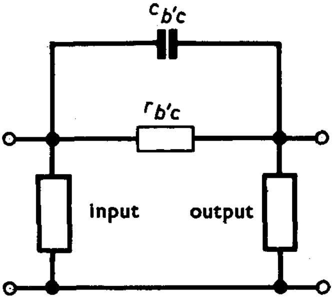

$c_{b'c}$
$r_{b'c}$
input
output

Fig. 364

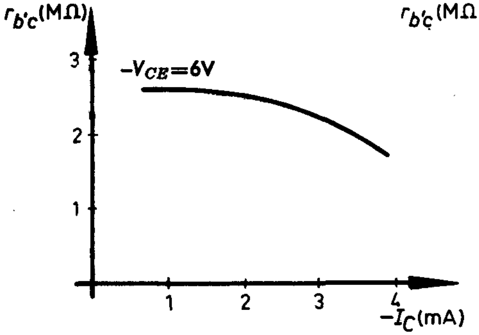


Fig. 365

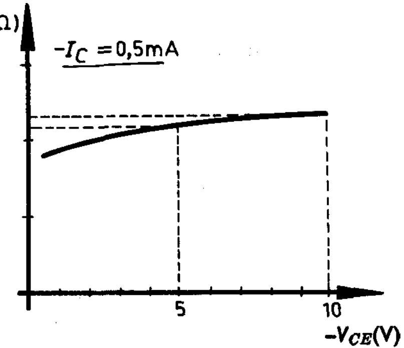


Fig. 366

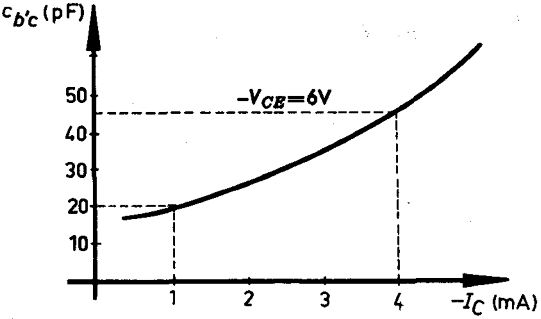


Fig. 367

The influence of the collector-emitter voltage on $c_{o\text{eq}}$

In Fig. 363 the capacitance $c_{o\text{eq}}$ is plotted as a function of the collector-emitter voltage. It is found that this capacitance decreases gradually as the collector current increases.

33.3. The feedback circuit

Fig. 364 represents the feedback circuit, consisting of the base-collector resistance $r_{b'c}$ with a capacitance $c_{b'c}$ connected parallel to it. We will now investigate the influence of the collector current and the collector-emitter voltage on the values of these two parameters.

The influence of the collector current on $r_{b'c}$

The base-emitter resistance $r_{b'c}$ is identical with the internal feedback-resistance of the transistor. The influence of the collector current on this resistance can be deduced from the transistor characteristics. This relationship is represented graphically in Fig. 365, which shows that this resistance decreases gradually as the collector current increases.

The influence of the collector-emitter voltage on $r_{b'c}$

The relationship between the base-collector capacitance $c_{b'c}$ and the collector current I_C is given in Fig. 367. This capacitance is found to increase fairly rapidly with the collector current, which effect must be ascribed to the increasing width of the base-collector junction as a result of which, of course, the base-collector capacitance increases.

The influence of the collector-emitter voltage on $c_{b'c}$

Fig. 368 represents the relationship between the base-collector capacitance $c_{b'c}$ and the base-emitter voltage $-V_{CE}$. We see from this curve that $c_{b'c}$ decreases as $-V_{CE}$ increases. Just as we did for the input and output circuits, we can represent the purely resistive effect and the purely capacitive effect of the feedback circuit (see Fig. 369) by the symbols $r_{r\text{eq}}$ and $c_{r\text{eq}}$ for this equivalent resistance and capacitance respectively. The values of $r_{r\text{eq}}$ and $c_{r\text{eq}}$ are expressed in terms of the transistor parameters $g_{bb'}$, $g_{b'c}$, $g_{b'e}$ and $c_{b'c}$ by means of the equations at the end of this chapter. We will now examine the influence of the collector current and of the collector-emitter voltage on this equivalent resistance and capacitance.

The influence of the collector current on $r_{r\text{eq}}$

The relationship between $r_{r\text{eq}}$ and $-I_C$ is plotted in Fig. 370. As the collector current increases, the value of $r_{r\text{eq}}$ also increases, which means that the internal feedback decreases. With the aid of this curve we can determine the value of $r_{r\text{eq}}$ as a function of the collector current.

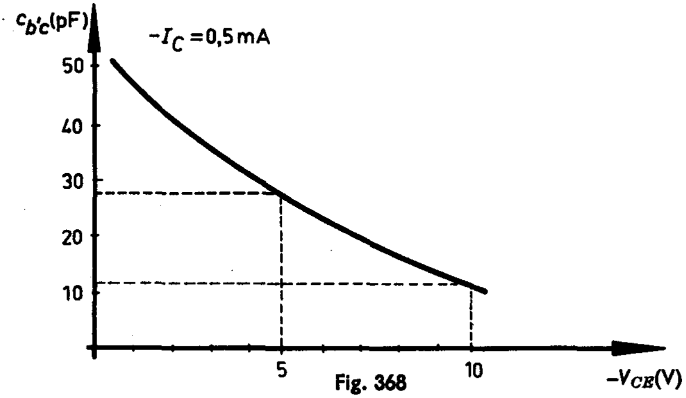
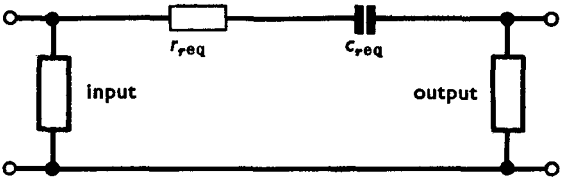


Fig. 368



r_{req}
c_{req}
input
output

Fig. 369

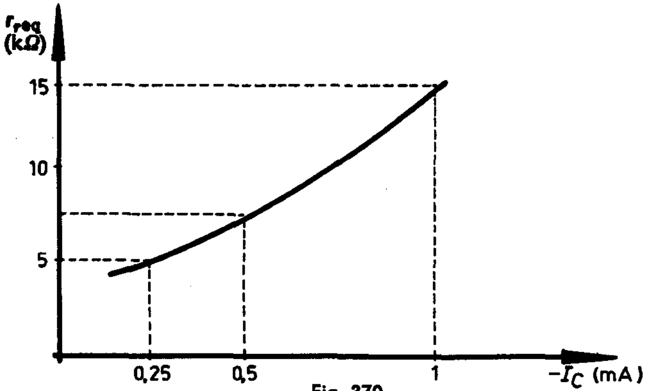


Fig. 370

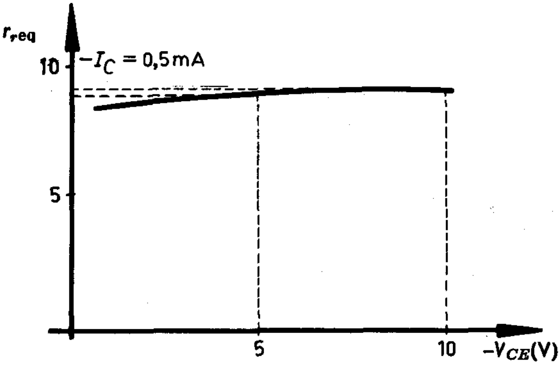


Fig. 371

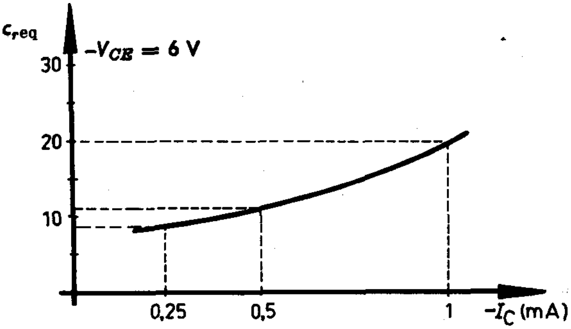


Fig. 372

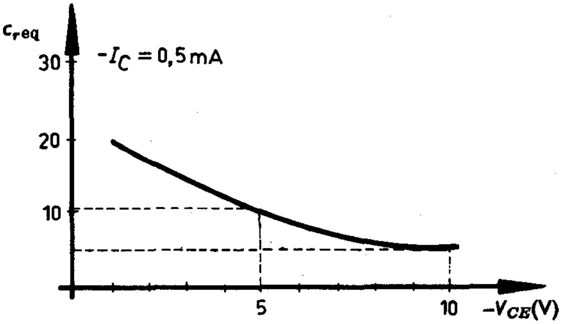


Fig. 373

The influence of the collector-emitter voltage on $r_{r\text{eq}}$

Fig. 371 shows $r_{r\text{eq}}$ plotted as a function of $-V_{CE}$. It is found that this voltage scarcely has any influence on the equivalent resistance $r_{r\text{eq}}$.

The influence of the collector current on $c_{r\text{eq}}$

Fig. 372 shows the relationship between $c_{r\text{eq}}$ and $-I_C$. In this case, the equivalent capacitance increases with the collector current. The shape of this curve can be deduced from the graph given in Fig. 367, in which the base-collector capacitance is plotted as a function of the collector current.

The influence of the collector-emitter voltage on $c_{r\text{eq}}$

The graph representing $c_{r\text{eq}}$ as a function of $-I_C$ (Fig. 373) can also be derived from the curve showing the relationship between the base-collector capacitance of the transistor and the collector-emitter voltage (Fig. 368).The capacitance decreases as the collector-emitter voltage increases.

33.4. Application of the curves

If we know the curves which represent the relationship between the various transistor parameters, we can accurately calculate the values of the various elements comprising the input, the output and the feedback circuit, for a given set of transistor operating conditions. In doing this, we must remember that in order to obtain the maximum gain at high frequencies – that is to obtain the highest possible values of the cut-off frequencies for the current gain and the slope – the operating conditions must be such that the constant collector current lies between 0.5 mA and 1 mA.

When transistors are employed at high frequencies they may be used under widely differing conditions. We will now examine various circuits and will investigate the behaviour of the transistor more closely in each of these.

Transistors can be used for the following purposes:

As amplifiers for wide frequency bands.
As aperiodic amplifiers.
As tuned amplifiers.

This shows that transistors can be used for quite different frequency ranges. Now the various transistor parameters shown in the equivalent circuit of Fig. 374 vary as a function of the frequency, and for this reason we will make a closer examination of the influence of the frequency of the input signal on the input circuit, the output circuit and the feedback circuit of the transistor.

Input

We can think of the input as consisting of a resistance $r_{i\mathrm{eq}}$ connected in parallel with a capacitance $c_{i\mathrm{eq}}$ (see Fig. 375), $r_{i\mathrm{eq}}$ being a pure resistance, and $c_{i\mathrm{eq}}$ being a pure capacitance. In fact the input of the transistor can be represented as a complex impedance, consisting of a real component and an imaginary component.

An impedance of this type can be divided into two parts, namely; a real component which can be represented by a pure resistance $r_{i\mathrm{eq}}$ and an imaginary component, which can be represented by a pure capacitance $c_{i\mathrm{eq}}$. Like the values of $r_{i\mathrm{eq}}$ and $c_{i\mathrm{eq}}$, the equivalent impedance of the circuit depends on the frequency of the input signal. The expressions for $r_{i\mathrm{eq}}$ and $c_{i\mathrm{eq}}$, which are given at the end of the chapter, also contain the angular frequency ω.

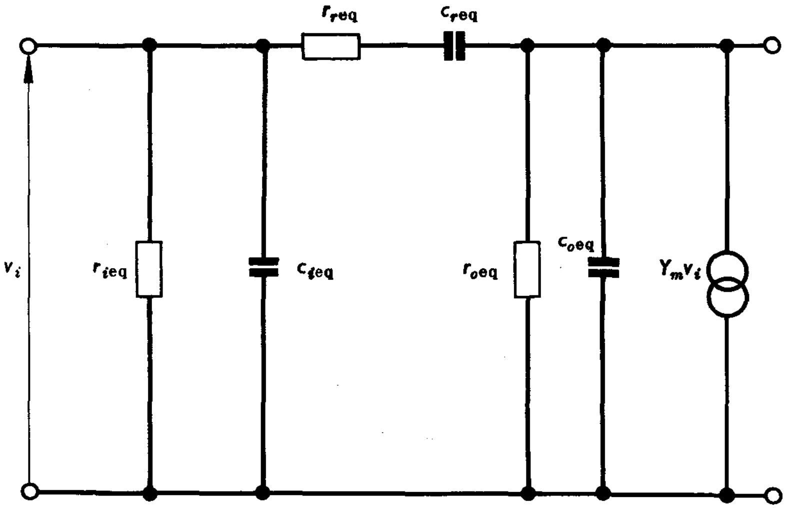

r_{req}
c_{req}
v_i
r_{ieq}
c_{ieq}
r_{oeq}
c_{oeq}
$Y_m v_i$

Fig. 374

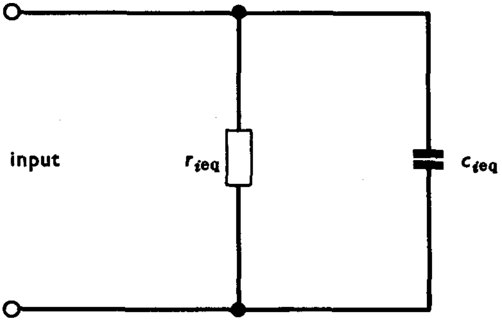

input
r_{ieq}
c_{ieq}

Fig. 375

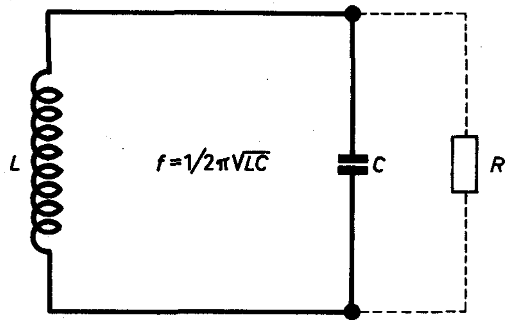

L
$f = 1/2\pi\sqrt{LC}$
C
R

Fig. 376

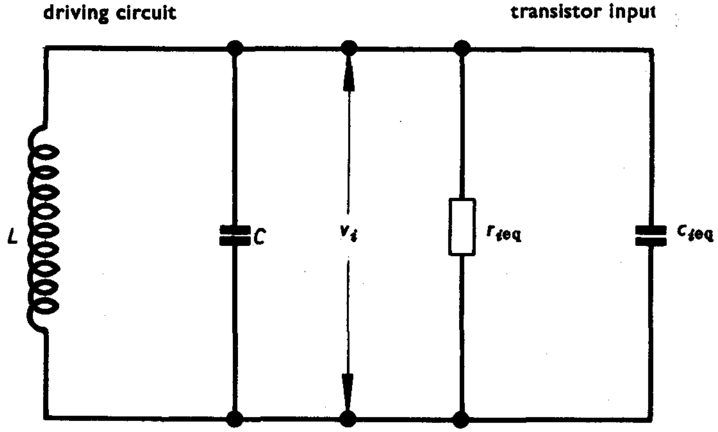

driving circuit
transistor input
L
C
v_i
r_{ieq}
C_{ieq}

Fig. 377

We will now consider a tuned circuit (Fig. 376). The resonent frequency of this circuit is determined by the values of the self inductance L and of the capacitance C. The damping is expressed by the value of the resistance R connected in parallel across the circuit. In a transistor circuit the input of the transistor is connected in parallel with the driving circuit which, for r.f. applications, is formed by a tuned circuit (see Fig. 377). Connecting the transistor across this circuit can thus be regarded as connecting a resistance $r_{i\mathrm{eq}}$ and a capacitance $c_{i\mathrm{eq}}$ across the circuit. The only result of connecting the resistance $r_{i\mathrm{eq}}$ will be that the damping of the tuned circuit increases, while the capacitance $c_{i\mathrm{eq}}$ will simply detune this circuit.

By considering the values of $r_{i\mathrm{eq}}$ and $c_{i\mathrm{eq}}$ as functions of the frequency, we can draw certain conclusions about the influence we may expect the transistor to have on the damping and on the detuning. These effects can be taken into account in determining the detuning curve of a receiver, and it may be possible to compensate for their influence on the sensitivity of the receiver, if this extends over a sufficiently large frequency range.

The influence of the frequency on r_{ieq}

We can measure the influence of the frequency of the input signal of a transistor on the value of r_{ieq} for a given value of a collector current. Fig. 378 shows such curves plotted for three different values of the collector current, i.e. for $-I_C = 1$ mA, 0.5 mA and 0.25 mA.

We see from this family of curves for a given transistor that r_{ieq} decreases sharply in the frequency range between 500 kc/s and a few Mc/s. At a frequency of 100 kc/s this resistance equals approximately 2.5 kΩ if the collector current is 1 mA and about 10 kΩ if this current is 0.25 mA. At a frequency of 1 Mc/s, these values are 1.5 kΩ and 6.5 kΩ respectively. This graph also shows that the smaller the collector current to which the transistor is adjusted, the more rapidly the resistance r_{ieq} decreases.

With the aid of such a graph, it is possible to determine the damping exercised by the transistor on the input circuit for various frequencies. For example, at a constant collector current of 0.5 mA and a frequency of 100 kc/s, the extra damping on this circuit will be approximately 5.0 kΩ (point *A*), at a frequency of 1 Mc/s approximately 2.5 kΩ (point *B*), and at a frequency of 3 Mc/s, about 2.0 kΩ (point *C*). The effects of this damping on the input circuit are represented for various frequencies in Figs. 379a, b and c.

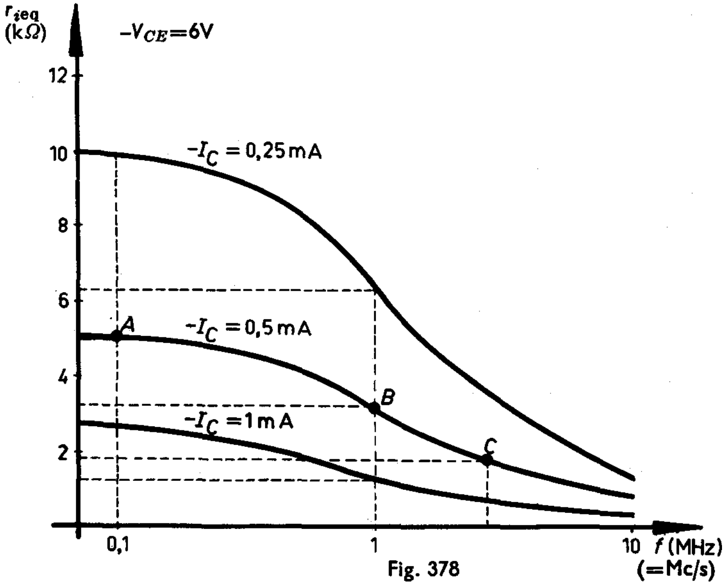


Fig. 378

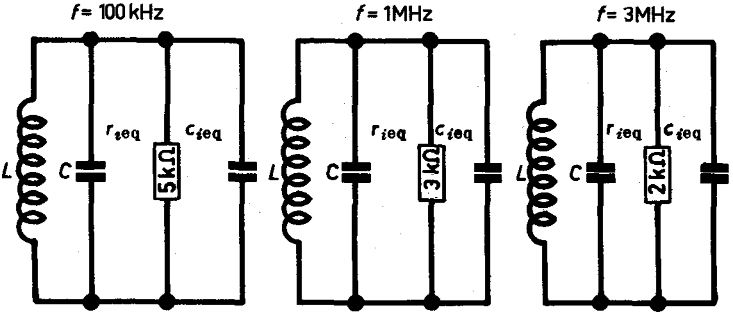

f= 100 kHz
f= 1MHz
f= 3MHz
L
C
r_{ieq}
c_{ieq}
5 kΩ
3 kΩ
2 kΩ

Fig. 379

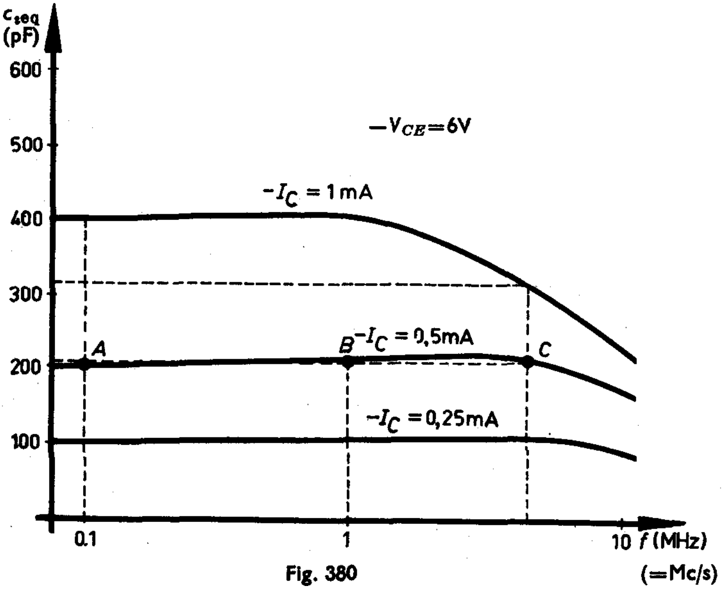
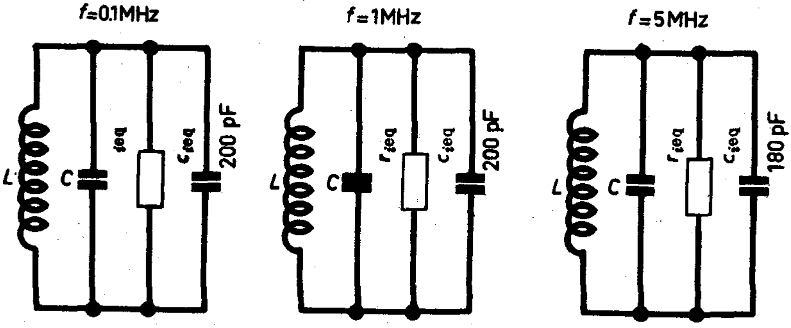


Fig. 380



f=0.1MHz
f=1MHz
f=5MHz
L
C
r_{ieq}
c_{ieq}
200 pF
200 pF
180 pF

Fig. 381

The influence of the frequency on c_{ieq}

Fig. 380 shows the equivalent capacitance c_{ieq} plotted as a function of the frequency for various values of the collector current ($-I_C = 1$ mA, 0.5 mA and 0.25 mA). This graph shows that c_{ieq} decreases with the collector current. At a frequency of 100 kc/s, c_{ieq} is about 400 pF if the collector current is adjusted to $-I_C = 1$ mA and about 100 pF if this current is adjusted to $-I_C = 0.25$ mA. At a frequency of 5 Mc/s, these values of c_{ieq} are 310 pF and 100 pF respectively for the same transistor operating conditions.

To start with, the input capacitance c_{ieq} remains practically constant and then drops gradually at higher frequencies. The larger the collector current, the sooner does this decrease occur. On the basis of such curves, it is possible to determine the influence of the input capacitance of the transistor on the detuning of the input circuit for various frequencies of the input signal. At a constant collector current $-I_C$ of 0.5 mA and a frequency of 100 kc/s, the value of c_{ieq} will be approximately 200 pF (point *A*) as it will at a frequency of 1 Mc/s (point *B*) and at a frequency of 5 Mc/s it will be about 180 pF. The effects of these changes in the input capacitance of the transistor on the input circuit are represented in Figs. 381 a, b and c.

The only effect which this capacitance has is to detune the input circuit. Since the variation of c_{ieq} with frequency is only slight at a constant collector current of 0.5 mA, this will not have any serious results. If the transistor had been adjusted to a constant collector current $-I_C$ of 1 mA, the effect of the variations in c_{ieq} would have been much greater.

Output

The output of the transistor consists of a pure resistance $r_{o\text{eq}}$ connected in parallel with a capacitance $c_{o\text{eq}}$ (see Fig. 382). As for the transistor input, the transistor output can be represented by a complex impedance, with a real component and an imaginary component. Such an impedance can be divided into two parts:

> A resistive component which can be represented by a pure resistance $r_{o\text{eq}}$.
>
> A reactive component which can be represented by a pure capacitance $c_{o\text{eq}}$.

Both of these elements can be expressed in terms of the various transistor parameters. The equivalent impedance of the circuit depends on the frequency of the input signal, as is also true for the real and imaginary components of the impedance, $r_{o\text{eq}}$ and $c_{o\text{eq}}$ respectively. Consequently the expressions for $r_{o\text{eq}}$ and $c_{o\text{eq}}$ which are given at the end of this chapter also contain the angular frequency ω.

The influence of the frequency on $r_{o\text{eq}}$

Fig. 383 shows $r_{o\text{eq}}$ plotted as a function of frequency with $-I_C$ as parameter. This graph shows that at a frequency of 100 kc/s the resistance $r_{o\text{eq}}$ is approximately 100 kΩ at a constant collector current of $-I_C = 0.25$ mA, and approximately 20 kΩ at a constant collector current of $-I_C = 1$ mA. At a frequency of 5 Mc/s, these values are 30 kΩ and 5 kΩ respectively. The smaller the value of the constant collector current $-I_C$ to which the transistor is adjusted, the more sharply does the value of $r_{o\text{eq}}$ decrease with increasing frequency.

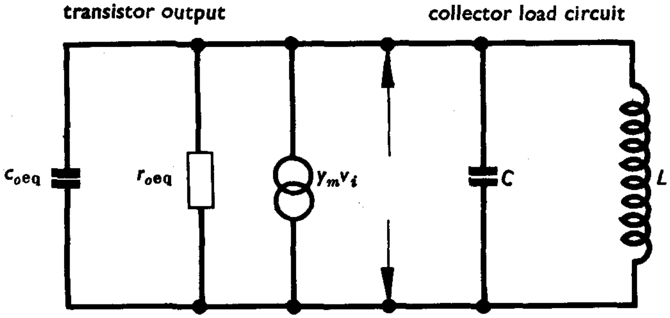

transistor output
collector load circuit
c_{oeq}
r_{oeq}
$y_m v_i$
C
L

Fig. 382

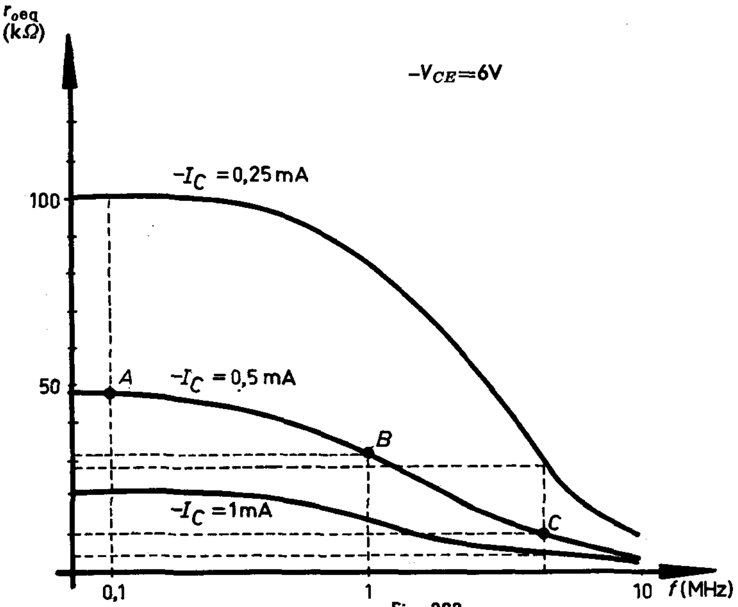


Fig. 383

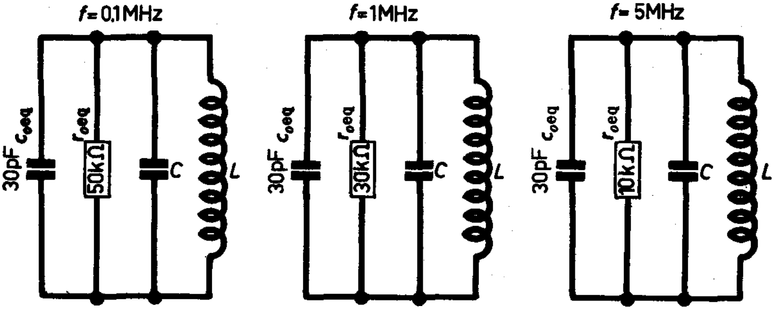

f=0,1MHz
30pF c_{oeq}
r_{oeq}
50kΩ
C
L
f=1MHz
30pF c_{oeq}
r_{oeq}
30kΩ
C
L
f=5MHz
30pF c_{oeq}
r_{oeq}
10kΩ
C
L

Fig. 384

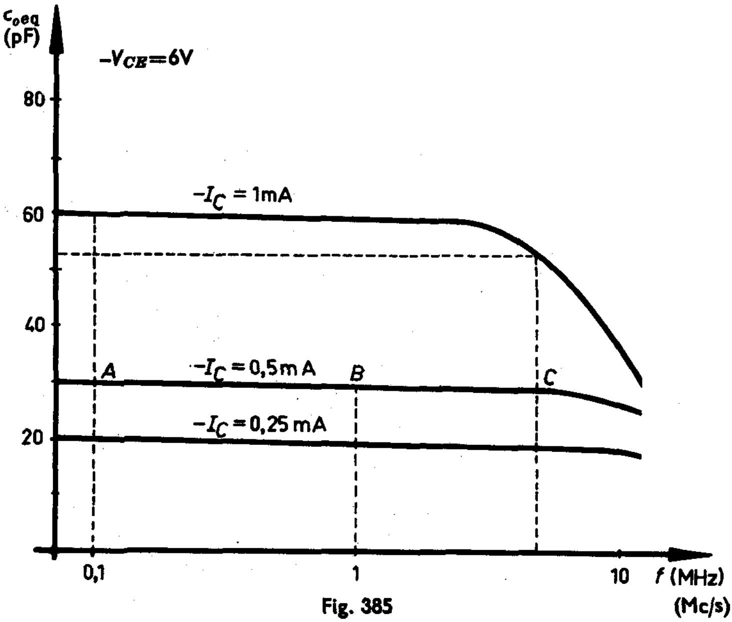


Fig. 385

At a constant collector current of $-I_C = 0.5$ mA (Fig. 383) r_{oeq} is 48 kΩ at a frequency of 100 kc/s, 35 kΩ at a frequency of 1 Mc/s and 10 kΩ at a frequency of 5 Mc/s. This shows that the extra damping exercised by the transistor on the output circuit decreases with increasing frequency. The effect of the transistor damping for various frequencies is represented in Figs. 384 a, b and c.

The influence of the frequency on c_{oeq}

The equivalent output capacitance c_{oeq} of the transistor is plotted in Fig. 385 as a function of frequency with the constant collector current $-I_C$ as parameter. At a frequency of 100 kc/s this capacitance is 60 pF at a constant collector current of 1 mA and 20 pF at a constant collector current of 0.25 mA. At a frequency of 5 Mc/s these values are 54 pF and 20 pF respectively.

Fig. 385 shows that at a low value of the constant collector current the output capacitance c_{oeq} is practically independent of the frequency. At higher values of this current the capacitance decreases at higher frequencies. The detuning caused by this capacitance if the transistor load consists of a tuned circuit is thus practically independent of the frequency of the input signal, provided that the transistor is adjusted to a very low value of $-I_C$.

Feedback

The feedback in the transistor can be represented by a resistance $r_{r\mathrm{eq}}$ connected in series with a capacitance $c_{r\mathrm{eq}}$ (see Fig. 386). Part of the output power is conveyed back to the input of the transistor via this circuit. The internal feedback depends on the values of $r_{r\mathrm{eq}}$ and $c_{r\mathrm{eq}}$.

The influence of the frequency on $r_{r\mathrm{eq}}$

Fig. 387 represents the relationship between the frequency and the value of $r_{r\mathrm{eq}}$ for various values of the collector current and shows that $r_{r\mathrm{eq}}$ decreases with increasing frequency. At a collector current of $-I_C = 0.5$ mA for example, $r_{r\mathrm{eq}}$ will be approximately 7 kΩ at a frequency of 100 kc/s, and approximately 4.5 kΩ at a frequency of 200 kc/s. At higher values of the frequency the curves run almost horizontally, showing that $r_{r\mathrm{eq}}$ does not drop any further. The graph also shows that the drop in $r_{r\mathrm{eq}}$ at relatively low frequencies increases with increasing value of the collector current.

The influence of the frequency on $c_{r\mathrm{eq}}$

It can be shown that the value of $c_{r\mathrm{eq}}$ is independent of the frequency. The expression given at the end of this chapter shows that $c_{r\mathrm{eq}}$ is determined exclusively by $g_{b'e}$, $g_{bb'}$, $g_{b'c}$ and $c_{b'e}$. It must not be forgotten, however, that the impedance which $c_{r\mathrm{eq}}$ represents for r.f. signals is inversely proportional to the frequency, as we see from the following equation.

$$Z_{cr} = 1/\omega c_{r\mathrm{eq}}.$$

As is shown in the following section, in order to neutralize the transistor we require to know the values of $r_{r\mathrm{eq}}$ and of $c_{r\mathrm{eq}}$. The value of R_3 can be read from the graph of Fig. 387.

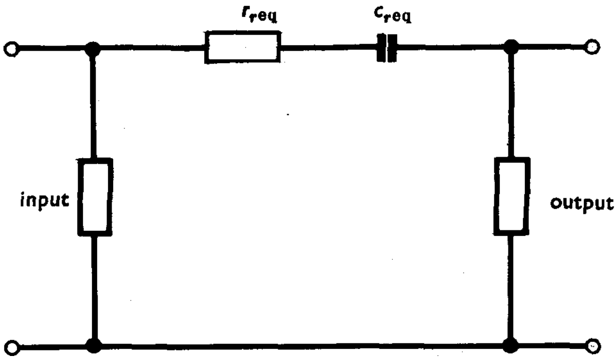

r_{req}
c_{req}
input
output

Fig. 386

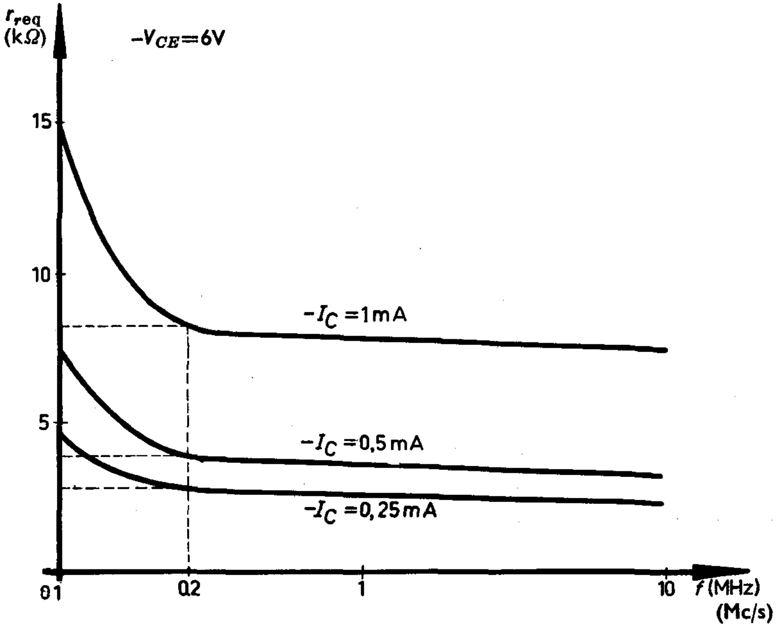


Fig. 387

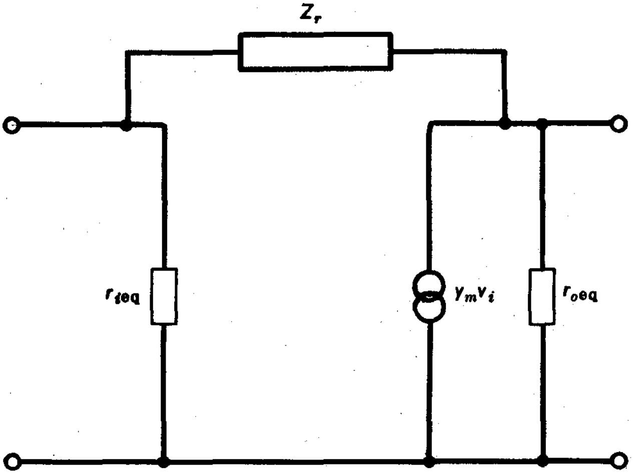

Z_r
r_{ieq}
$y_m v_i$
r_{oeq}

Fig. 388

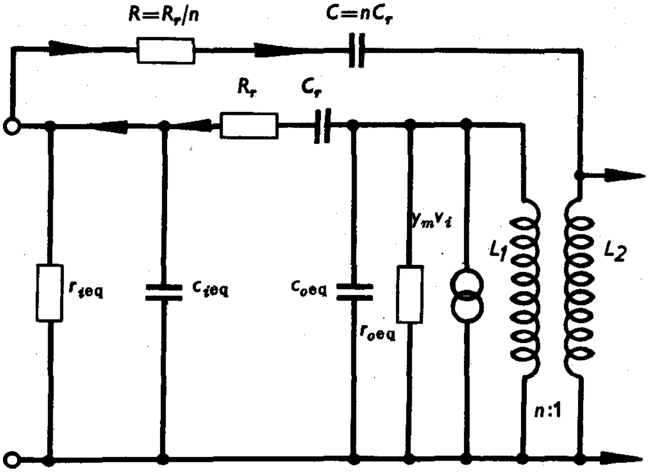

$R=R_r/n$
$C=nC_r$
R_r
C_r
r_{ieq}
c_{ieq}
c_{oeq}
r_{oeq}
$y_m v_i$
L_1
L_2
$n:1$

Fig. 389

33.5. Neutralization

From the circuit of Fig. 388 we see that part of the voltage across the equivalent output impedance $r_{o\text{eq}}$ is returned to the equivalent input impedance $r_{i\text{eq}}$ of the transistor by its internal feedback, which we have represented by the impedance Z_r.

This effect can be cancelled out (neutralized) by simultaneously conveying to the input of the transistor a voltage of the same amplitude but of opposite phase as the feedback voltage. The principal of this form of neutralization of the transistor is represented in Fig. 389. The output voltage is present across the terminals of coil L_1 (the primary of a transformer) which is connected across the output of the transistor. Part of this voltage is returned via R_rC_r to the terminals of $r_{i\text{eq}}$.

Across the secondary of this transformer (L_2) there is a voltage of opposite phase to the primary voltage; the amplitude of this voltage is determined by the transformation ratio between L_2 and L_1:

$$n = V_{L1}/V_{L2},$$

where V_{L1} represents the voltage across the primary and V_{L2} that across the secondary. If $n = 1$ the voltage across L_2 will thus be equal to, but of opposite phase to that across L_1 but in phase opposition to the latter.

In order to compensate the voltage which is fed back to the input via the circuit R_rC_r, all that is necessary is to connect L_2 to the input of the transistor via a resistance $R = R_r$ in series with a capacitance $C = C_r$. If the transformation ratio n is not equal to 1, for example $n = 5$, the voltage available across the terminals of L_2 will only be 1/5th of the voltage across the terminals of L_1. In order to supply the input of the transistor with a voltage of the same amplitude as that which must be ascribed to the circuit R_rC_r, we must now reduce R to $R_r/n = R_r/5$ and increase C to $nC_r = 5C_r$. Whatever the value of n, neutralization will have to be carried out by applying a resistance in series with a capacitance, the value of the resistance being given by $R = R_r/n$ and that of the capacitance by $C = nC_r$.

33.6. Maximum power gain

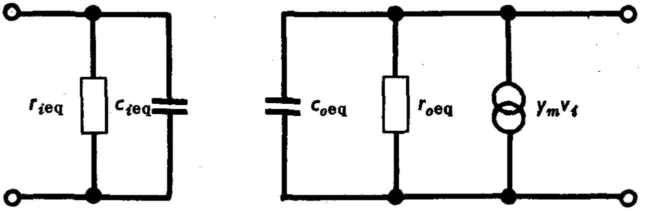
If the internal feedback is cancelled out by neutralization we can calculate the maximum value of the power gain of a transistor in an r.f. circuit. To do this, we shall start from the equivalent circuit of Fig. 390. The values of $r_{i\mathrm{eq}}$, $c_{i\mathrm{eq}}$, $r_{o\mathrm{eq}}$, $c_{o\mathrm{eq}}$ and Y_n can be determined for each value of the frequency of the input signal. The quantities $c_{i\mathrm{eq}}$ and $c_{o\mathrm{eq}}$ represent the purely capacitive effect of the input and output of the transistor respectively. These quantities only affect the tuning of the input and output circuits. In order to determine the input power and the available output power we only have to take account of the quantities $r_{i\mathrm{eq}}$ and $r_{o\mathrm{eq}}$. The maximum power gain is obtained when the load impedance R_L is equal to $r_{o\mathrm{eq}}$. Fig. 391 represents the equivalent circuit for this special case.

The power gain is given by the quotient of the power available at the output and the power conveyed to the input:

$$G_P = P_o/P_i.$$

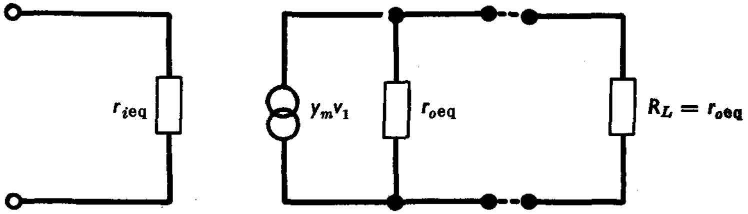
In Fig. 392 this maximum power gain is plotted as a function of the frequency of the input signal. We see from this graph that the maximum power gain of a transistor rapidly decreases with increasing frequency, once the latter has exceeded a certain value.

At a frequency of 100 kc/s (point *A*) the gain is 43 dB, as also at a frequency of 500 kc/s (point *B*); at a frequency of 1 Mc/s the gain is still 40 dB (point *C*), but at a frequency of 5 Mc/s it has dropped to 20 dB (point *D*). The shape of this curve must be taken into account in designing an r.f. amplifier.



r_{ieq}
c_{ieq}
c_{oeq}
r_{oeq}
$y_m v_1$

Fig. 390



r_{ieq}
$y_m v_1$
r_{oeq}
$R_L = r_{oeq}$

Fig. 391

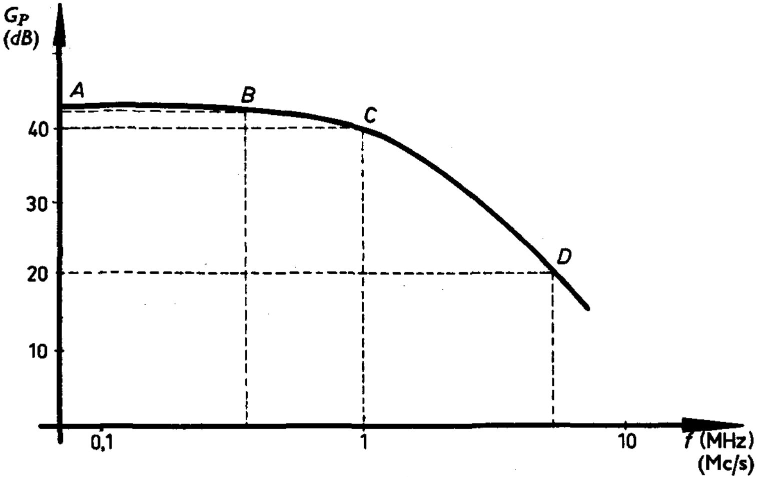


Fig. 392

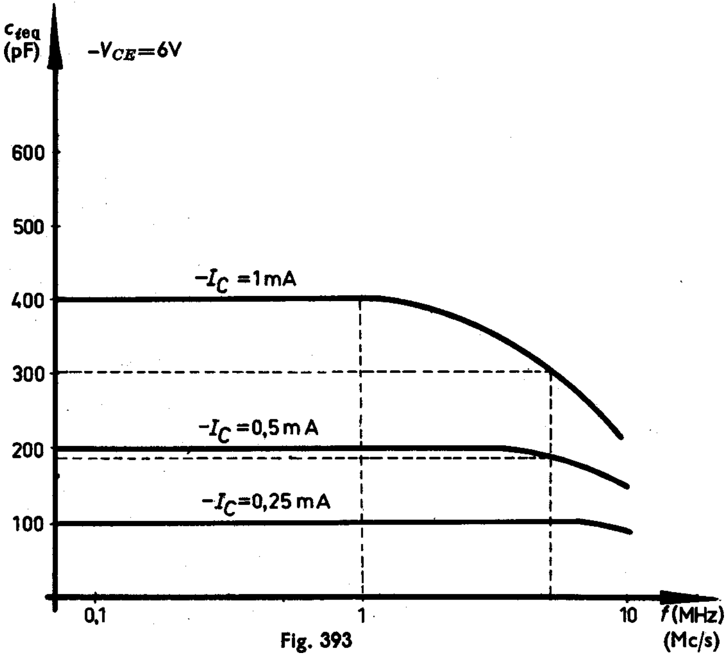

c_{feq} (pF)
$-V_{CE}=6V$
600
500
400
300
200
100
$-I_C=1 mA$
$-I_C=0,5 mA$
$-I_C=0,25 mA$
0,1
1
10
f (MHz) (Mc/s)

Fig. 393

33.7. The effect of the dependence of the transistor parameters on current, voltage and frequency

Knowing that the transistor parameters depend on the current, voltage and frequency as just described, we can investigate the behaviour of a transistor for all r.f. applications. In this field of application the transistor can be used as a wide-band amplifier, as a selective amplifier for a narrow band (i.f. amplifier) or as an amplifier for the normal r.f. range (in r.f. stages & mixer stages). We will now examine the influence of the transistor input and output for each of these applications.

Amplifier for a wide frequency band

An amplifier for a wide frequency band will have a low load impedance so that there is no need to apply neutralization.

The influence of the transistor input

The input capacitance of the transistor will exercise a very strong influence on the behaviour of the input circuit in relation to the input signal. The curves of Fig. 393 show that the lower the value of $-I_C$ to which the transistor is adjusted, the less will this capacitance depend on the frequency.

The influence of the transistor output

As the output impedance of the transistor will always be made considerably higher than the load impedance, the damping of the latter by the output impedance can usually be neglected. The load impedance will usually be of the order of 1 kΩ, while the output impedance will be of the order of some tens of kΩ. The variations of the output capacitance, which is in fact relatively low, may usually be neglected.

Selective amplifier for a narrow frequency band

In a selective amplifier the load impedance will normally have a high value, being approximately equal to the output impedance of the transistor, so that neutralization has to be applied. As a rule the load impedance will consist of a tuned circuit, while in most cases the input circuit will be aperiodic; at a frequency of 450 kc/s for example, the relatively low value of the input impedance of the transistor (of the order of 1 kΩ) would damp a tuned input circuit to an unacceptable extent, unless the transistor was connected to a tap on the latter. The influence of the frequency on the various transistor parameters need not be considered here since the amplifier always works at a given frequency, i.e. the intermediate frequency of the receiver.

The influence of the transistor input

At the input of the amplifier we can distinguish a real component and an imaginary component, $r_{i\mathrm{eq}}$ and $c_{i\mathrm{eq}}$ respectively in Figs. 394 and 395. The real component is responsible for the damping which the transistor input exercises on the input circuit, and the imaginary component is responsible for the detuning of the input circuit. If the input circuit is not tuned the influence of the transistor input capacitance can be left out of consideration.

The input impedance is thus the principal factor involved here; by determining this, it is possible to select the optimum matching, that is, with the impedance of the input circuit being equal or practically equal to the transistor input impedance.

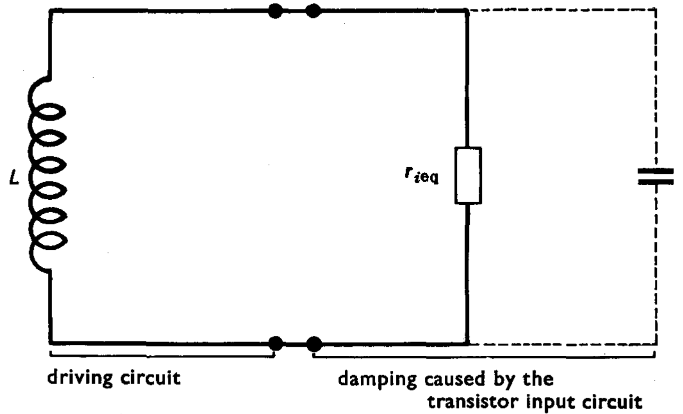

L
r_{ieq}
driving circuit
damping caused by the transistor input circuit

Fig. 394

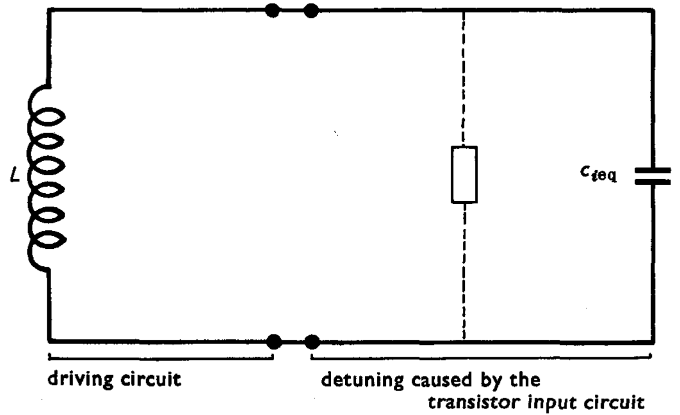

L
C_{ieq}
driving circuit
detuning caused by the transistor input circuit

Fig. 395

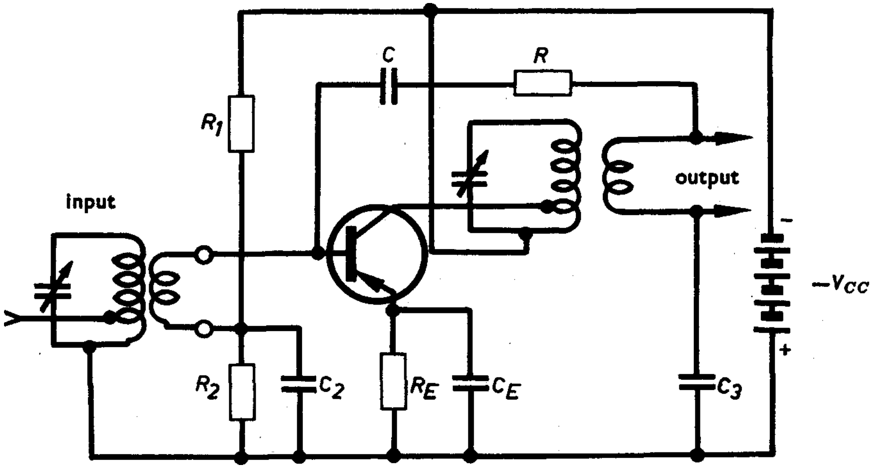

C
R
R1
input
output
-
-VCC
+
R2
C2
RE
CE
C3

Fig. 396

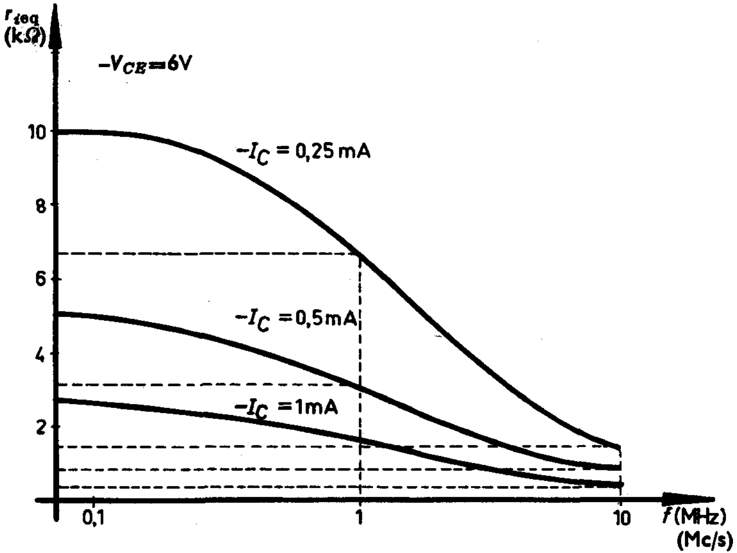


Fig. 397

The influence of the transistor output

The transistor output impedance is usually lower than the impedance of the output circuit. In order to prevent the transistor output impedance having too great a damping effect on the output circuit, it is usual to employ a transformer with a tapping on the primary as indicated in Fig. 396. In this way, the impedance of that part of the output circuit which now constitutes the transistor load can be made practically equal to the transistor impedance. The damping resulting from this output impedance will thus be relatively slight. For a given value of $-I_C$ the detuning caused by the transistor output capacitance can be read from the curve $c_{oeq} = f(-I_C)$ on page 379.

R.F. amplifier for a wide frequency band

If a transistor is used in an r.f. amplifier or the mixer stage of a radio receiver it will be required to give the most constant amplification possible over a fairly wide frequency range. It is therefore very important to take into account the influence of the frequency of the input signal on the various transistor parameters.

The influence of the transistor input

The maximum gain is obtained when the impedance of the input circuit is equal to the input impedance of the transistor. However, as the latter varies with the frequency according to the characteristic curves shown in Fig. 397, matched drive will only be possible for one given frequency. Nevertheless, if the constant collector current $-I_C$ is made sufficiently low, the value of r_{ieq} will only vary slightly, so that the resulting amplification will remain constant within narrow limits. At a low value of the constant collector current, the influence of the frequency on the input capacitance can be completely neglected.

The influence of the transistor output

The output impedance of the transistor decreases with increasing frequency as shown in Fig. 398, and since this impedance is connected in parallel across the output circuit the damping which it produces will increase with the frequency of the signal to be amplified. The lower the value of the collector current to which the transistor is adjusted, the smaller will this damping be. From Fig. 399, in which the output capacitance $c_{o\text{eq}}$ is plotted as a function of the frequency with $-I_C$ as parameter, we see that this capacitance remains practically constant, whatever the frequency of the input signal, provided that the constant collector current is less than 1 mA.

The damping of the input and output circuits finds expression in the curve which represents the power gain as a function of the frequency (see page 403).

Feedback

In a selective amplifier the feedback between the output and input of the transistor stage may be left out of consideration if it has been neutralized. In the equivalent circuit diagrams the feedback is represented by a resistance $r_{r\text{eq}}$ in series with a capacitance $c_{r\text{eq}}$. Now Fig. 400 shows that this resistance $r_{r\text{eq}}$ decreases with increasing frequency. If $r_{r\text{eq}}$ decreases, the feedback from the output circuit to the input circuit will increase, and will therefore change the power gain of the stage. In order to limit this change to the minimum, the constant collector current must be adjusted to the lowest possible value.

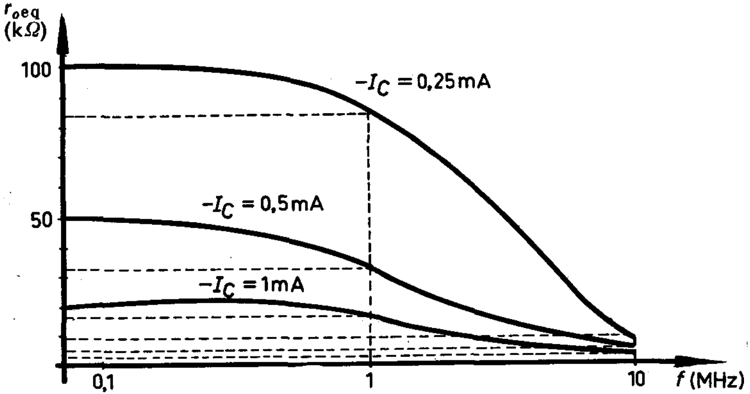


Fig. 398

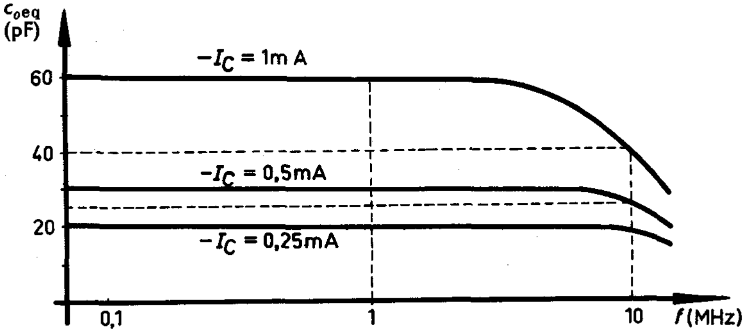


Fig. 399

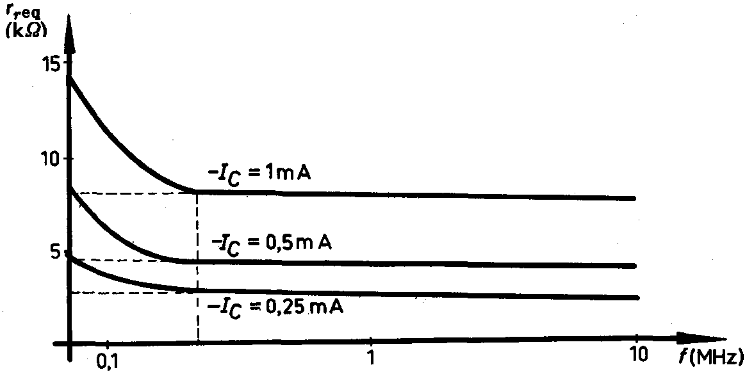


Fig. 400

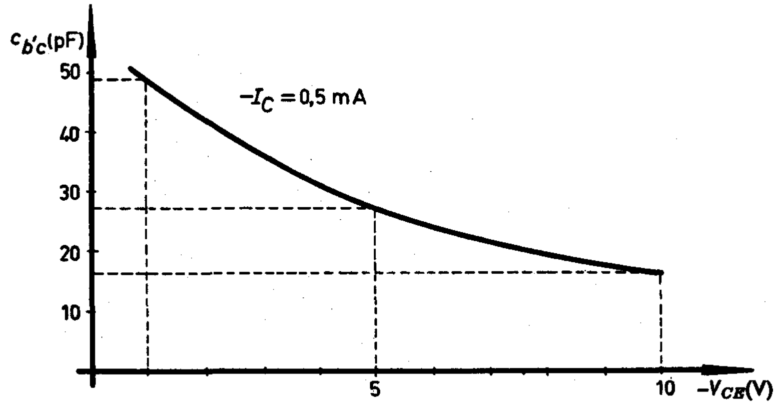


Fig. 401

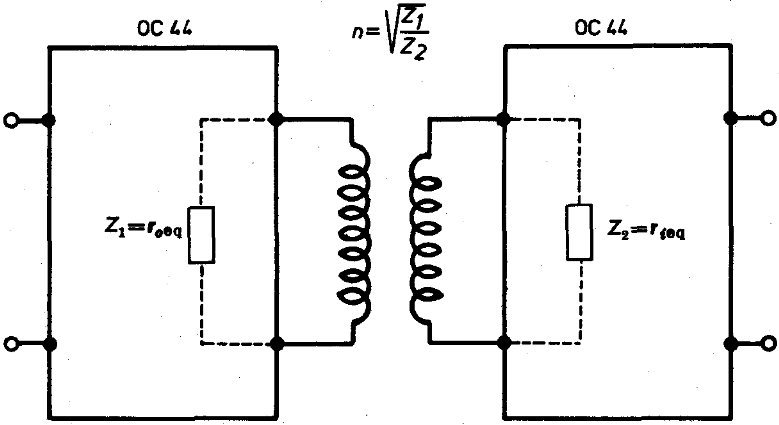

OC 44
$n=\sqrt{\frac{Z_1}{Z_2}}$
OC 44
$Z_1=r_{oeq}$
$Z_2=r_{ieq}$

Fig. 402

The curve which represents the influence of the collector-emitter voltage on the base-collector capacitance $c_{b'e}$ (Fig. 401), shows that this capacitance increases as the collector-emitter voltage decreases. If there is a decrease in the voltage of the battery supplying the receiver, $c_{b'e}$ may increase to such an extent that the neutralization becomes insufficient. In this case a capacitive feedback may develop between the output and input of the transistor stage and may even lead to oscillation.

The influence of the frequency on the optimum matching

Fig. 402 is the equivalent circuit of an r.f. amplifier stage employing an OC 44 transistor. The influence of the frequency on the input impedance of the transistor was discussed on page 390. At a frequency of 500 kc/s this impedance $r_{i\text{eq}}$ is approximately 4.5 kΩ and at 1.5 Mc/s its value is about 1.5 kΩ. The influence of the frequency on the output impedance was discussed on page 394; at a frequency of 500 kc/s this impedance is about 40 kΩ and at 1.5 Mc/s approximately 18 kΩ.

The load of this stage consists of the input impedance of the following stage, in this case the mixer stage. Suppose that the same type of transistor (OC 44) is used in the mixer stage, adjusted to a collector current of 0.3 mA. At a frequency of 500 kc/s its input impedance will be about 7.5 kΩ. At 500 kc/s the two stages will be matched if the transformation ratio n (see Fig. 402), is:

$$n = \sqrt{Z_1/Z_2} = \sqrt{40/7.5} \backsimeq 2.3.$$

At a frequency of 1.5 Mc/s the transformation ratio required for correct matching will be:

$$n = \sqrt{Z_1/Z_2} = \sqrt{18/3.6} \backsimeq 2.23.$$

This shows that the reduction of the output impedance of the first stage and of the input impedance of the second stage which occur with increasing frequency, will have little affect on the optimum matching. If the frequency increases from 500 kc/s to 1.5 Mc/s, the transformation ratio for optimum matching only decreases from 2.3 to 2.23 (see Fig. 403).

33.8. Expressions for the circuit elements occurring in the equivalent circuit diagrams

A transistor can be represented by the equivalent circuit of Fig. 404. In this chapter however, in our investigation of the influence of the frequency on the behaviour of a transistor, we have employed the circuit of Fig. 404, which includes the following circuit elements:

$$r_{i\mathrm{eq}},\ r_{o\mathrm{eq}},\ c_{i\mathrm{eq}},\ c_{o\mathrm{eq}},\ y_m,\ r_{r\mathrm{eq}} \text{ and } c_{r\mathrm{eq}}.$$

It may be of interest to express these elements in terms of the various transistor parameters and of the frequency.

$$r_{i\mathrm{eq}} = \frac{1}{g_{bb'}} + \frac{g_{bb'} + g_{b'e}}{g_{b'e}g_{bb'} + g_{b'e}^2 + \omega^2 (c_{b'e} + c_{b'c})^2},$$

$$g_{o\mathrm{eq}} = \frac{1}{r_{o\mathrm{eq}}} = g_{ce} + g_{b'c} + g_m \cdot \frac{g_{b'c}(g_{bb'} + g_{b'e})\,\omega^2 c_{b'c}c_{b'e},}{(g_{bb'} + g_{b'e})^2 + \omega^2 c_{b'e}^2},$$

$$c_{i\mathrm{eq}} = \frac{c_{b'e} + c_{b'c}}{(1 + g_{b'e}/g_{bb'})^2 + \omega^2(c_{b'e} + c_{b'c})^2/g_{bb'}^{2}},$$

$$c_{o\mathrm{eq}} = c_{b'c}\left\{1 + \frac{g_m(g_{bb'} + g_{b'e})}{(g_{bb'} + g_{b'e})^2 + \omega^2 c_{b'e}^2}\right\},$$

$$|y_m| = \frac{g_m}{(1 + g_{b'e}/g_{bb'})^2 + \omega^2(c_{b'e} + c_{b'c})^2/g_{bb'}^{2}},$$

$$r_{r\mathrm{eq}} = \frac{1}{g_{bb'}}(1 + c_{b'e}/c_{b'c}) + \frac{g_{b'c}(1 + g_{b'e}/g_{bb'})}{\omega^2 c_{b'c}^2},$$

$$c_{r\mathrm{eq}} = \frac{c_{b'c}}{(1 + g_{b'e}/g_{bb'}) - c_{b'e}g_{b'c}/c_{b'c}g_{bb'}}.$$

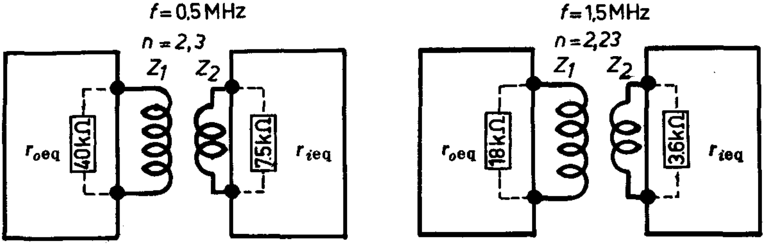

f = 0,5 MHz
n = 2,3
Z1
Z2
$r_o$eq
40kΩ
75kΩ
$r_i$eq
f = 1,5 MHz
n = 2,23
Z1
Z2
$r_o$eq
18kΩ
3,6kΩ
$r_i$eq

Fig. 403

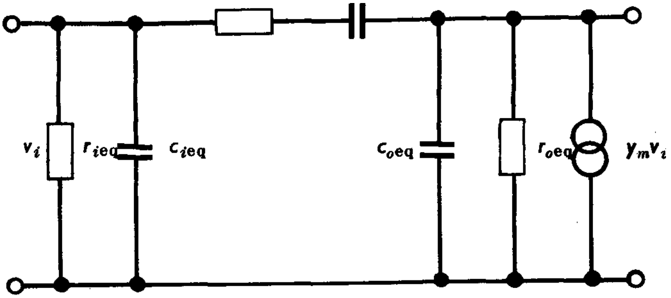

v_i
$r_i$eq
$c_i$eq
$c_o$eq
$r_o$eq
$y_m v_i$

Fig. 404

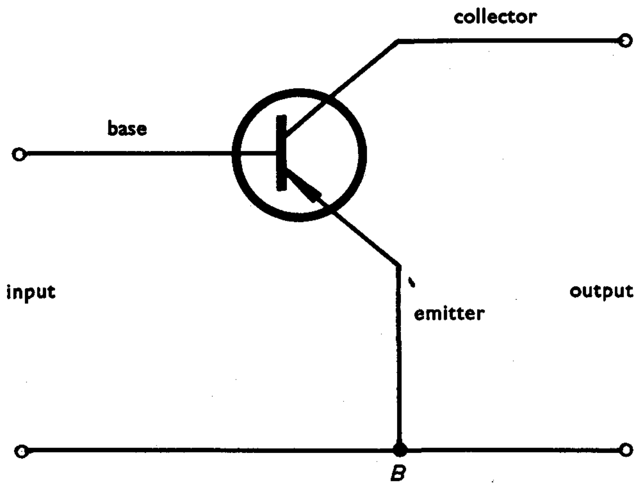

collector
base
input
emitter
output
B

Fig. 406

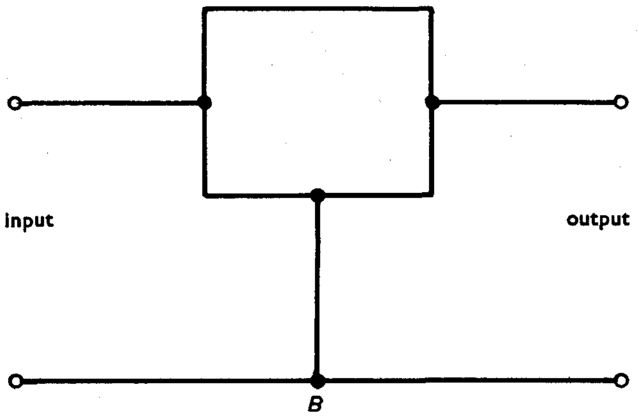

input
output
B

Fig. 407

CHAPTER 34

Possible circuit configurations for transistors

If a transistor (see Fig. 406) is used in an amplification stage it can be represented as forming part of an input circuit on one hand and of an output circuit on the other hand (Fig. 407). In this figure the input and output circuits have the point *B* in common. One of the transistor electrodes is connected to this point and thus forms the common point of the input and output circuits.
The transistor can be connected in three ways:

In the common emitter configuration.
In the common base configuration.
In the common collector configuration.

The parameters by means of which the operation of the transistor can be described differ according to the configuration which is employed. A closer examination of the problem enables us to follow the variations of these parameters, and from them to draw conclusions concerning the possible applications of the different configurations.
The common emitter configuration is by far the most frequently used; we shall see that this configuration does in fact offer two important advantages over the two other configurations, namely:

A much greater power gain.
The smallest possible difference between the input and output impedances, which simplifies the application of this circuit.

The transistor manufacturer publishes characteristics for the common base configuration and for the common emitter configuration. It is also possible to define the operation of the transistor in the common collector configuration.

CHAPTER 35

The common-emitter configuration

When a transistor is used in common emitter (Fig. 408), the generator is connected in the input circuit between the base and the emitter, and the output R_L is connected in the output circuit between the collector and emitter of the transistor. The input and output circuits thus have the emitter in common.

In the equivalent circuit of Fig. 409, we see that the signal to be amplified is applied between the base and the emitter, while the output signal is taken off between the collector and the emitter.

The operation of a transistor is determined by the three following important factors:

The current gain of the stage.
The input impedance and its variations.
The output impedance and its variations.

35.1. Power gain

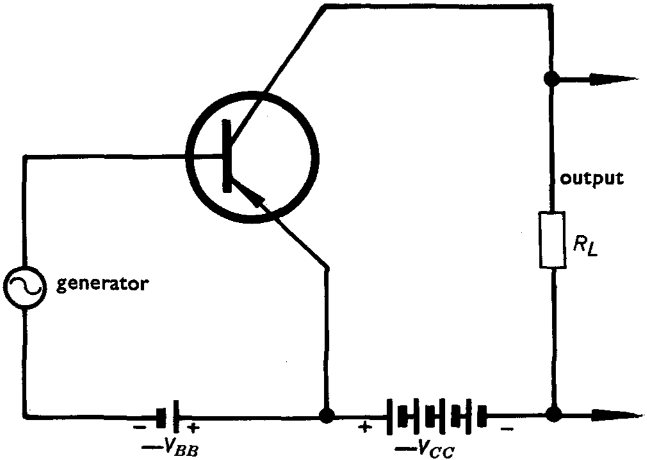
The power gain is given by the quotient of the power available at the output and the power conveyed to the input:

$$G_P = P_o/P_i.$$

As was shown on page 330, the power gain equals the product of the voltage gain and the current gain. The voltage gain is equal to the quotient of the output voltage and the input voltage, and the current gain equals the quotient of the output current and the input current.

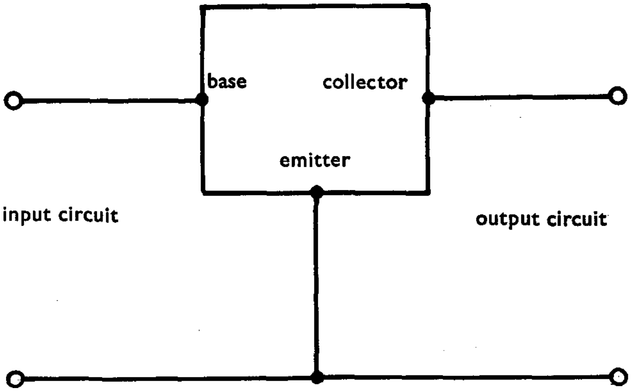
Voltage gain

The voltage gain is given by the voltage across the load impedance v_{RL} and the voltage v_g supplied by the generator. In common emitter the voltage gain is always greater than one and is sometimes very large. The voltage gain which is obtained depends on the value of the load resistance R_L.



output
R_L
generator
– +
$-V_{BB}$
+ –
$-V_{CC}$

Fig. 408



base
collector
emitter
input circuit
output circuit

Fig. 409

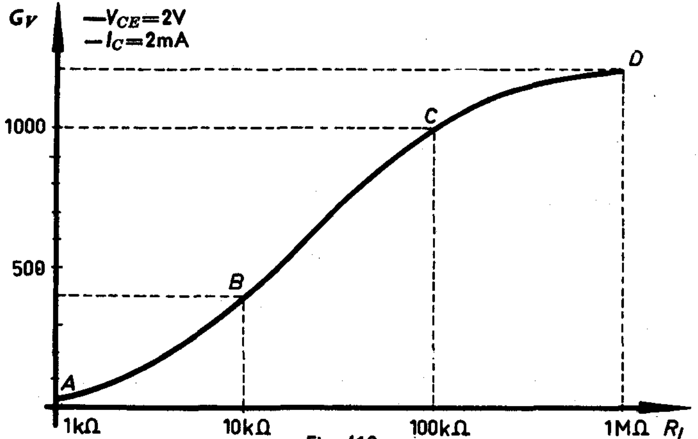


Fig. 410

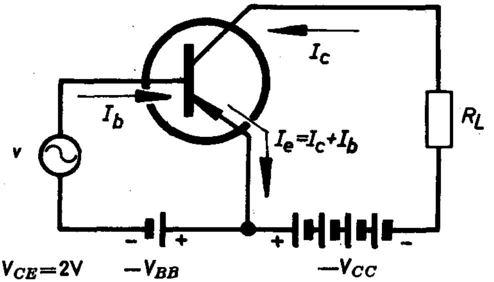

I_c
I_b
$I_e=I_c+I_b$
R_L
v
−
+
$-V_{BB}$
+
−
$-V_{CC}$

Fig. 411

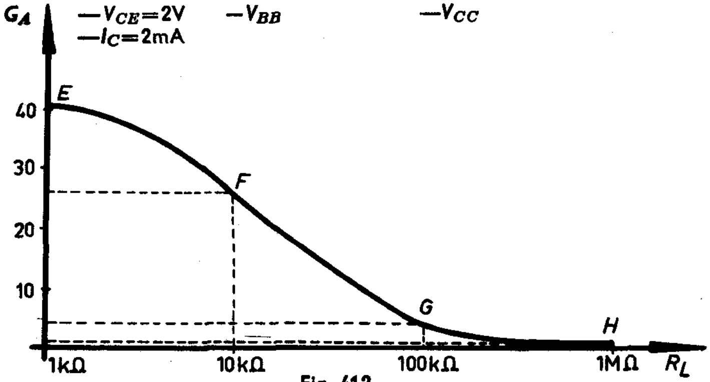


Fig. 412

The influence of R_L on the voltage gain is represented graphically in Fig. 410. This curve has already been discussed in detail on page 333. In common emitter the voltage gain is found to increase with the value of R_L. For example, it is 50 at $R_L = 1\ k\Omega$ (point *A*), 400 at $R_L = 10\ k\Omega$ (point *B*), 1000 at $R_L = 100\ k\Omega$ (point *C*), and 1200 at $R_L = 1\ M\Omega$ (point *D*). In practice the voltage gain of a transistor in common emitter with a high value of R_L will lie between 500 and 1.000.

Current gain

The current gain is given by the quotient of the output current and the input current. As we see from Fig. 411, the output current is equal to the collector current i_c, while the input current in common emitter is equal to the base current. Since this collector current is always greater than the base current, the current gain of a transistor in common emitter will also always be greater than one.

The size of the current gain depends above all upon the value of the load resistance R_L in the output circuit. In Fig. 412 the current gain is plotted as a function of R_L; this curve has already been discussed on page 334. For example, the current gain is

> 40 at $R_L = 1\ k\Omega$ (point *E*), 25 at $R_L = 10\ k\Omega$ (point *F*), 5 at $R_L = 100\ k\Omega$ (point *G*) and 0.8 at $R_L = 1\ M\Omega$ (point *H*).

In practice, the current gain of a transistor in common emitter will usually lie between 40 and 100 if R_L has a low value.

Conclusion

If the current gain and the voltage gain are known for a given value of the load resistance, the power gain can be calculated from the product of these two values.

$$G_P = G_V G_A.$$

From the curves representing the voltage gain and the current gain as functions of R_L, it is possible to calculate the form of the curve representing the power gain as a function of R_L (see Fig. 413). At $R_L = 1\ \text{k}\Omega$ the voltage gain is 50 (point *A*) and the current gain is 40 (point *E*) so that the power gain is:

$$G_P = G_V G_A = 50 \times 40 = 2000 = 33.0\ \text{dB (point } J\text{)}.$$

At $R_L = 10\ \text{k}\Omega$ the voltage gain is 400 (point *B*) and the current gain is 25 (point *F*), so that the power gain is:

$$G_P = G_V G_A = 400 \times 25 = 10\,000 = 40\ \text{dB (point } K\text{)}.$$

In the same way for $R_L = 100\ \text{k}\Omega$, we find that $G_V = 1000$ (point *C*) and $G_A = 5$ (point *G*), so that:

$$G_P = G_V G_A = 1000 \times 5 = 5\,000 = 37\ \text{dB (point } L\text{)}.$$

At $R_L = 1\ \text{M}\Omega$, we have $G_V = 1\,200$ (point *D*) and $G_A = 1$ (point *H*), so that:

$$G_P = G_V G_A = 1\,200 \times 1 = 1\,200 = 30.8\ \text{dB (point } M\text{)}.$$

The curve obtained in this way shows that the power gain is a maximum when the load impedance is equal to the output impedance of the transistor; in the case under consideration this is at

$$R_L = h_o = 20\ \text{k}\Omega.$$

In all practical applications a transistor in common emitter will have a current gain and a voltage gain both greater than 1. The power gain, the product of these two quantities, can thus reach a very high value.

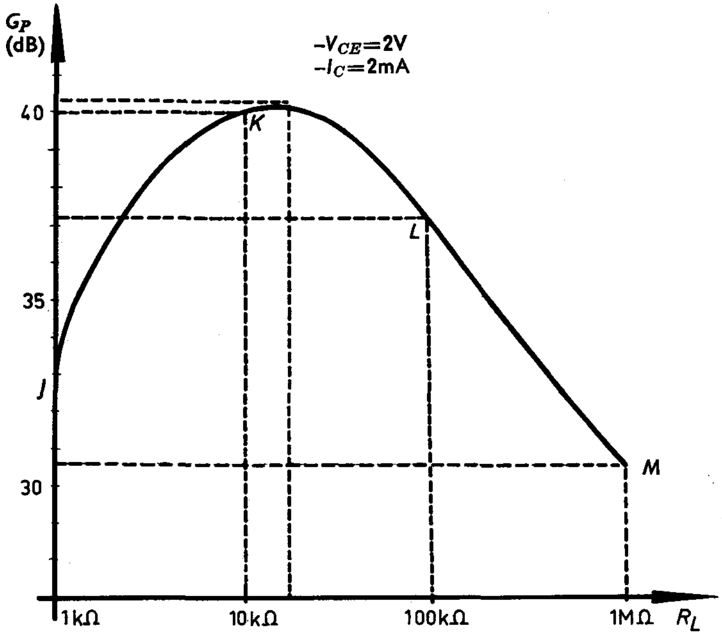


Fig. 413

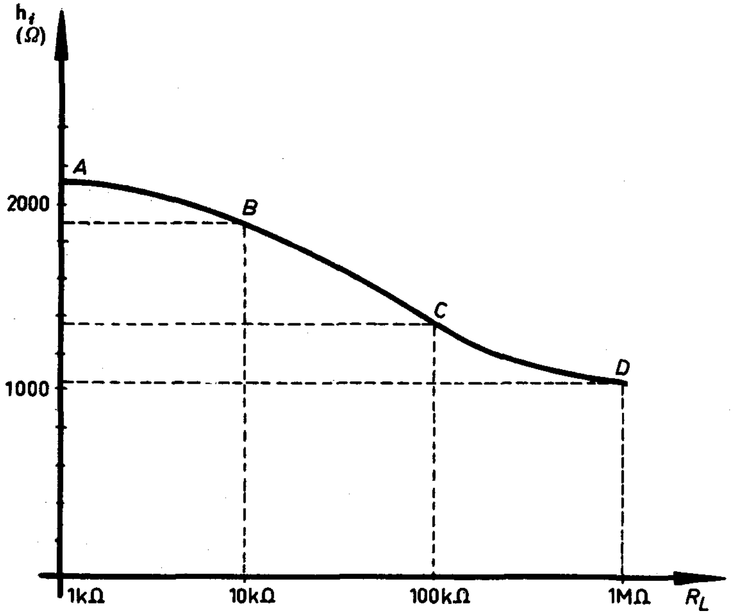


Fig. 414

35.2. The influence of the load impedance on the input impedance of the transistor

The influence of the collector current on the input impedance of a transistor has already been discussed. This impedance also varies with the value of the load impedance. In order to plot the input impedance as a function of the load impedance we must keep the collector-emitter voltage $-V_{CE}$ and the collector current $-I_C$ constant.

The curve $h_i = f(R_L)$ obtained in this way is given in Fig. 414, which shows that the input impedance decreases slightly as the load impedance R_L increases. At $R_L = 1$ kΩ the input impedance of the transistor in question (OC 71) was 2.1 kΩ (point A), at $R_L = 10$ kΩ this impedance was 1900 Ω (point B), at $R_L = 100$ kΩ it was 1.4 kΩ (point C) and at $R_L = 1$ MΩ it has dropped to 1.1 kΩ (point D). These measurements were carried out at $-V_{CE} = 2$ V and $-I_C = 0.5$ mA. Such curves can also be obtained for other values of the collector current, and it is found that for increasing values of $-I_C$ the curve shifts in the direction of the abscissa.

The average value of the input impedance of a transistor in common emitter is of the order of 1 kΩ at $-I_C = 2$ mA. It is particularly important to know this value in order to obtain an insight into the difference between the output impedance and the input impedance of a transistor; the smaller this difference is, the easier it is to obtain good matching between two transistors connected in the same configuration.

35.3. The influence of the generator impedance on the output impedance of the transistor

The output impedance of a transistor decreases with increasing collector current but is also dependent on the generator impedance. In this case too, the collector-emitter voltage $-V_{CE}$ and the collector current $-I_C$ must be kept constant when we plot the curve representing the relationship between the generator impedance R_g and the output impedance of the transistor.

Fig. 415 represents the curve obtained in this way. We see that the output impedance only decreases slightly when the generator impedance increases. For example, at $R_g = 10\ \Omega$ the output impedance is $\mathrm{h}_o = 22\ \mathrm{k}\Omega$ (point *A*), at $R_g = 100\ \Omega$ the value is $\mathrm{h}_o = 20\ \mathrm{k}\Omega$ (point *B*), at $R_g = 1\ \mathrm{k}\Omega$ it is $\mathrm{h}_o = 16\ \mathrm{k}\Omega$ (point *C*), and at $R_g = 10\ \mathrm{k}\Omega$ we have $\mathrm{h}_o = 14\ \mathrm{k}\Omega$ (point *D*). Thus we see that a change in the generator impedance in the ratio of 1 to 1,000, that is from 10 Ω to 10 kΩ, results in the output impedance changing in the ratio of 1 to 0.6, i.e. from 22 kΩ to 14 kΩ.

The curve of Fig. 415 applies to a constant collector current $-I_C$ of 2 mA; at a higher value of this current the output impedance of a transistor decreases, corresponding to a shift of the curve in the direction of the abscissa.

At a constant collector current of $-I_C = 2$ mA the input impedance of the transistor is found to be of the order of 1 kΩ, while the output impedance is of the order of 16 kΩ. The difference between the output impedance and the input impedance of a transistor in common emitter is thus relatively small. Good matching is easy to achieve; with transformer coupling, the required turns ratio is given by $n = (\mathrm{h}_o/\mathrm{h}_i)^{\frac{1}{2}}$.

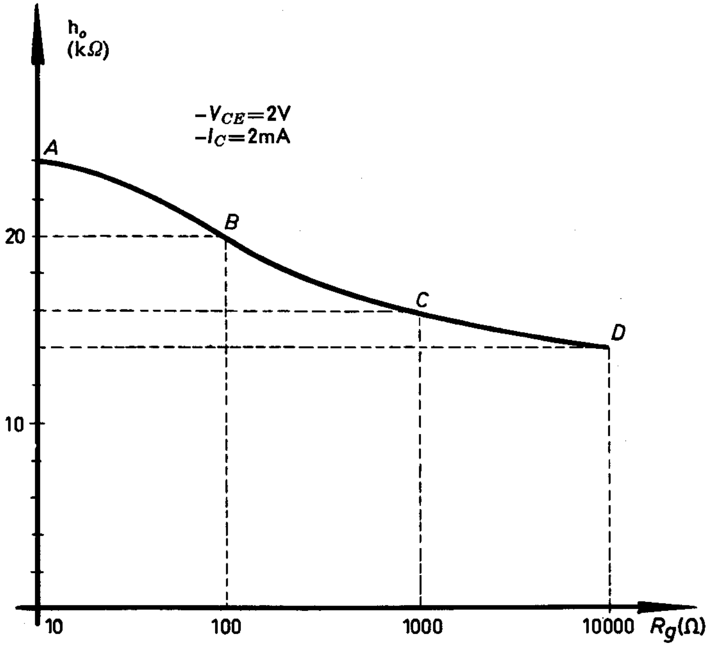


Fig. 415

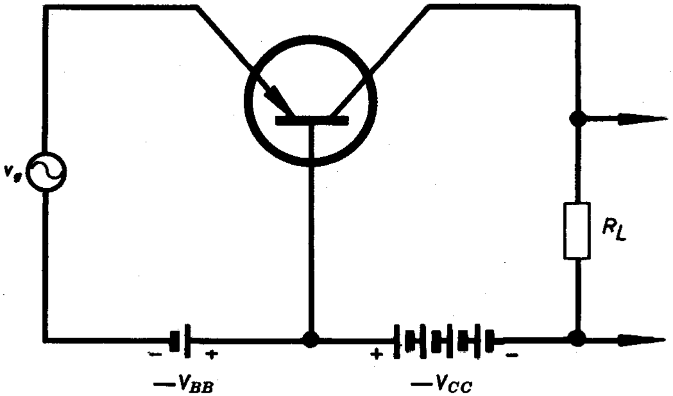

v_g
R_L
− +
—V_BB
+ −
—V_CC

Fig. 416

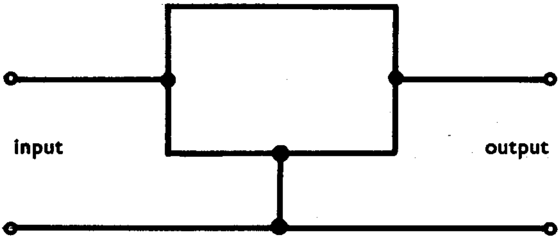

input
output

Fig. 417

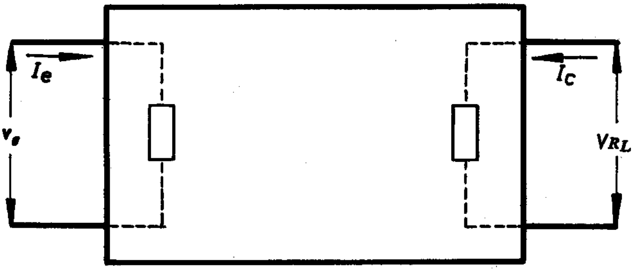


Fig. 418

CHAPTER 36

The common-base configuration

In the common-base configuration the generator is connected between the base and the emitter of the transistor, while the signal is taken off across the load impedance between the collector and the base (see Fig. 416). This corresponds to the equivalent circuit of Fig. 417, in which the input circuit is connected between the emitter and the base, and the output circuit between the collector and the base. The input and output circuits thus have the base in common. As for the previous circuit, we shall discuss the influence of the load impedance on the power gain, and that of the load impedance and of the generator impedance on the input impedance and the output impedance of the transistor respectively.

36.1. Power gain

In this case too, the power gain of the circuit equals the product of the voltage gain and the current gain of the stage:

$$G_P = G_V G_A.$$

In this expression the voltage gain is given once more by the quotient of the output voltage and the input voltage, while the current gain is given by the quotient of the output current and the input current (see Fig. 418).

Voltage gain

The voltage gain is given by the quotient of the voltage V_{RL} across the load impedance and the voltage v_g delivered by the generator according to the expression:

$$G_V = v_o/v_g = v_{RL}/v_g.$$

In the common base configuration the voltage gain is always greater than 1 and is sometimes very large; this voltage gain is practically the same as the gain attained in common emitter.

The voltage gain depends on the load impedance R_L, as will be seen from the curve in Fig. 419. At $R_L = 1\ \mathrm{k}\Omega$ we have $G_V = 40$ (point *A*), at $R_L = 10\ \mathrm{k}\Omega$ the value is $G_V = 380$ (point *B*) at $R_L = 100\ \mathrm{k}\Omega$ it is $G_V = 950$ (point *C*) and at $R_L = 1\ \mathrm{M}\Omega$ the value is $G_V = 1150$ (point *D*).

The voltage gain thus increases with the load impedance and the increase proves to be very considerable in the range between a few kilohms and 100 kilohms. At low values of R_L the voltage gain approaches a minimum value, while at very high values of R_L it approaches a maximum value. In common base the voltage gain will usually be between 500 and 1,000.

Current gain

The current gain is determined by the ratio of the output current to the input current. From Fig. 420 we see that the output current corresponds to the collector current of the transistor and the input current corresponds to the emitter current.

The relationship between the collector current and the emitter current is given by:

$$I_e = I_c + I_b,$$

from which it follows that the collector current is

$$I_c = I_e - I_b.$$

The collector current is thus smaller than the emitter current. This means that the output current of a transistor in common base will be smaller than the input current, so that the current gain will always be less than 1.

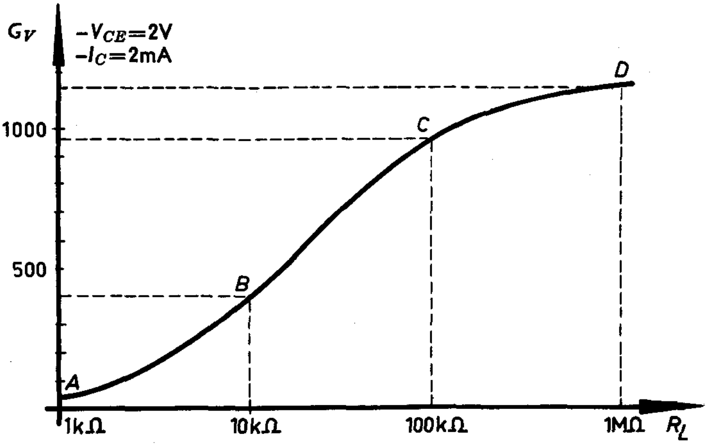


Fig. 419

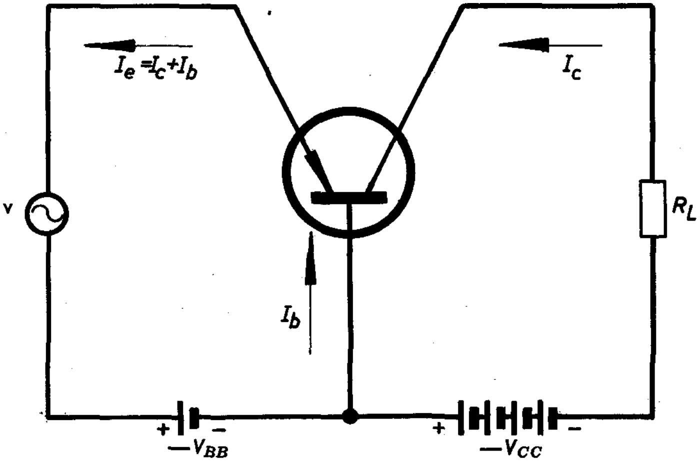

$I_e = I_c + I_b$
I_c
v
R_L
I_b
+
−
$-V_{BB}$
+
−
$-V_{CC}$

Fig. 420

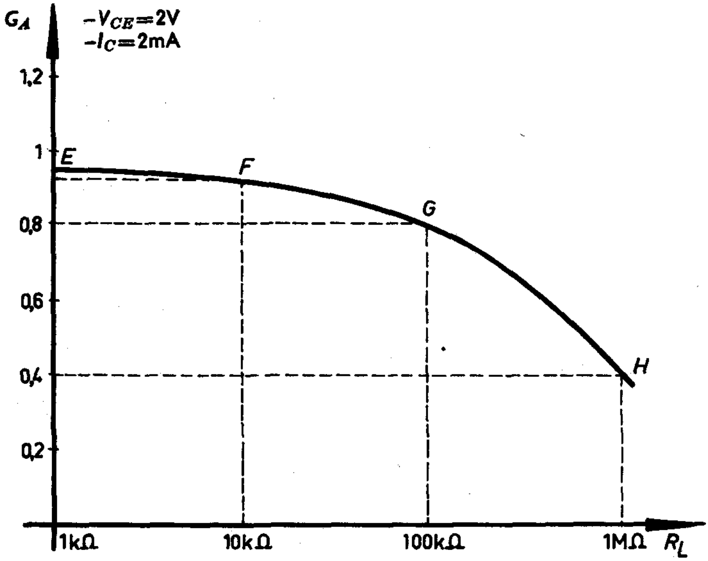


Fig. 421

The current gain of a transistor in common base is dependent on the load impedance. The influence of this impedance R_L on the current gain G_A is represented graphically in Fig. 421.
At $R_L = 1$ kΩ we have $G_A = 0.95$ (point *E*), at $R_L = 10$ kΩ the value of the current gain is $G_A = 0.92$ (point *F*), at $R_L = 100$ kΩ it is $G_A = 0.8$ (point *G*), and at $R_L = 1$ MΩ the value is $G_A = 0.4$ (point *H*).

The current gain of a transistor in common base thus decreases at the load impedance increases; in the range between $R_L = 1$ kΩ and $R_L = 100$ kΩ the influence of the load impedance is only slight, but it increases sharply in the range between $R_L = 100$ kΩ and $R_L = 1$ MΩ. The current gain will always be less than 1, whatever the value of the load impedance.

Conclusion

As already mentioned, the power gain is given by the product of the voltage gain G_V and the current gain G_A. With the aid of these curves, which show the voltage and current gains as functions of the load impedance, we can determine the power gain for various values of the load impedance.

The voltage gain of a transistor in common base is practically equal to that of a transistor in common emitter, but the current gain is always much lower (always less than one). The power gain of a transistor in the former configuration is smaller than the gain which can be obtained in common emitter.

The voltage gain G_V is shown as a function of R_L in Fig. 422 and the current gain G_A is given as a function of the same quantity in Fig. 423. The power gain of the transistor, as obtained from these two curves, is plotted in Fig. 424 as a function of R_L. At $R_L = 1\ \text{k}\Omega$ we have $G_V = 80$ (point *A*) and $G_A = 0.95$ (point *E*), from which follows the power gain:

$$G_P = G_V G_A = 80 \times 0.95 = 76 = 18.8 \text{ dB (point } J\text{)}.$$

At $R_L = 10\ \text{k}\Omega$ the corresponding values are $G_V = 400$ (point *B*) and $G_A = 0.92$ (point *F*), from which it follows that the power gain is:

$$G_P = G_V G_A = 400 \times 0.92 = 368 = 25.7 \text{ dB (point } K\text{)}.$$

At $R_L = 100\ \text{k}\Omega$ we have $G_V = 950$ (point *C*) and $G_A = 0.8$ (point *G*), from which the power gain is:

$$G_P = G_V G_A = 950 \times 0.8 = 760 = 28.8 \text{ dB (point } L\text{)}.$$

At $R_L = 1\ \text{M}\Omega$ the values are $G_V = 1150$ (point *D*) and $G_A = 0.4$ (point *H*), from which it follows that the power gain is:

$$G_P = G_V G_A = 1150 \times 0.4 = 460 = 26.6 \text{ dB (point } M\text{)}.$$

The curve obtained in this way shows that the power gain is a maximum at $R_L \backsimeq 180\ \text{k}\Omega$. The output impedance of a transistor in common base is of the order of several hundred kilohms.

At lower values of the load impedance R_L the power gain drops rapidly, but at higher values of this impedance the decrease is less rapid.

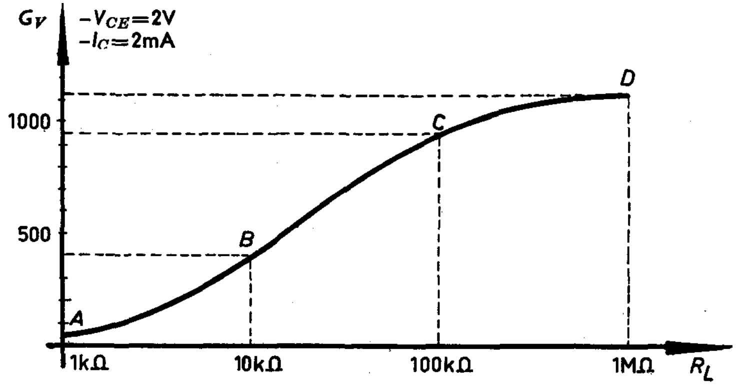


Fig. 422

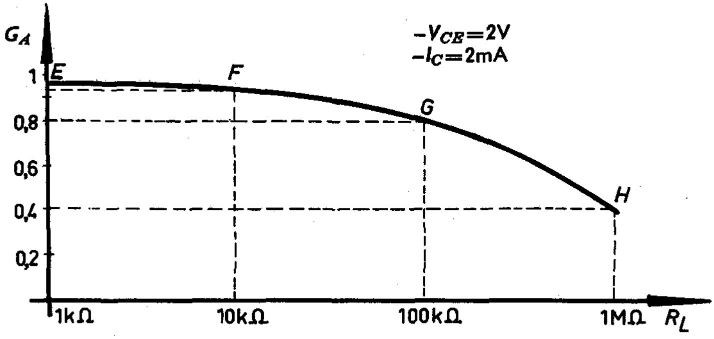


Fig. 423

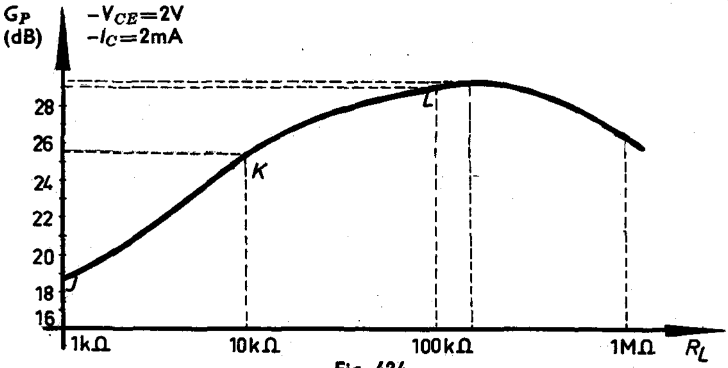


Fig. 424

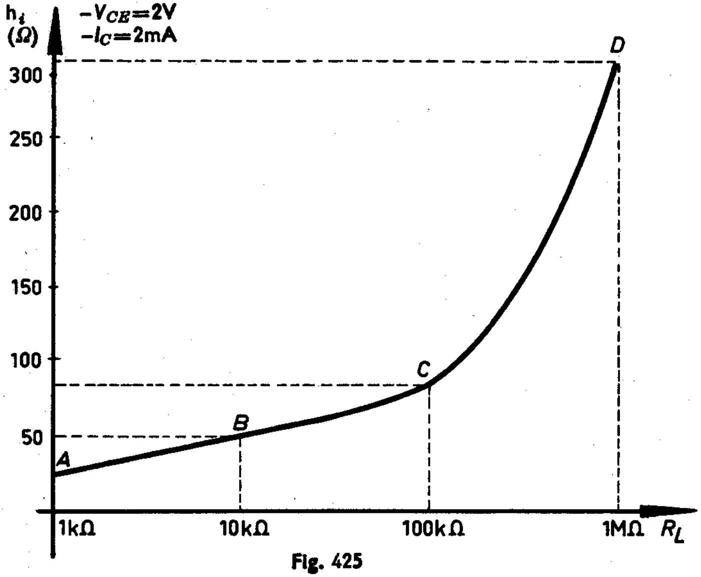

h_i
(Ω)
$-V_{CE}=2V$
$-I_C=2mA$
300
250
200
150
100
50
A
B
C
D
1kΩ
10kΩ
100kΩ
1MΩ
R_L

Fig. 425

36.2. The influence of the load impedance on the input impedance of the transistor

The input impedance of a transistor in common base decreases as the collector current increases. It is also important to know the influence of the load impedance on this input impedance. This relationship is plotted in Fig. 425 for constant values of the collector- emitter voltage $-V_{CE}$ and collector current $-I_C$. At $R_L = 1$ kΩ the input impedance h_i equals 25 Ω (point *A*), at $R_L = 10$ kΩ it is $h_i = 50$ Ω (point *B*), at $R_L = 100$ kΩ we have $h_i = 80$ Ω (point *C*) and at $R_L = 1$ MΩ the value is $h_i = 300$ Ω (point *D*).

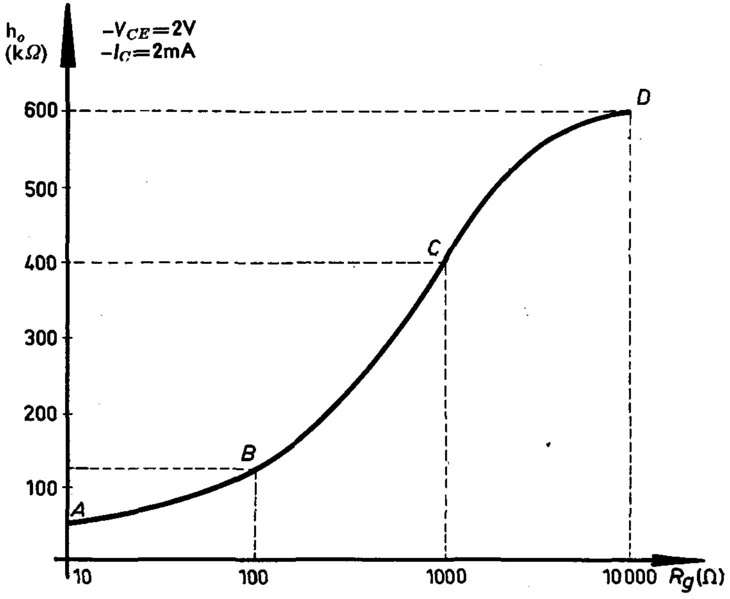
We see from this that the input impedance increases rapidly at high values of the load impedance (of the order of megohms). Curves of this type can be plotted for various values of the collector current. The curve of Fig. 425 refers to a collector current $-I_C$ of 2 mA; at higher values of the collector current the curve shifts in the direction of the abscissa. The input impedance is noticeably lower than in common emitter, and will usually be about 50 Ω.

36.3. The influence of the generator impedance on the output impedance of the transistor

The output impedance of a transistor in common base decreases as the collector current increases. It is also important to know the influence of the generator impedance on the output impedance. Fig. 426 shows the output impedance h_o plotted as a function of the generator impedance R_g. At $R_g = 10\ \Omega$ we have $h_o = 50$ kΩ (point *A*), at $R_g = 100\ \Omega$ the corresponding value is $h_o = 125$ kΩ (point *B*) at $R_g = 1$ kΩ it is $h_o = 400$ kΩ (point *C*) and at $R_g = 10$ kΩ we have $h_o = 600$ kΩ (point *D*). This curve shows that the output impedance increases rapidly with increasing generator impedance; in particular h_o increases very rapidly in the range between $R_g = 100\ \Omega$ and $R_g = 3$ kΩ, i.e. from approximately 125 kΩ to 550 kΩ. At low and very high values of R_g the influence of the generator impedance on the output impedance of the transistor is much less.

It is also possible to plot the relationship between R_g and h_o for other values of the collector current than $-I_C = 2$ mA; it is found that the curve of Fig. 426 then shifts in the direction of the abscissa.

The output impedance of a transistor in common base will usually be of the order of several hundred kilohms. As the input impedance, by contrast, is only about 50 Ω, it is difficult to match two amplifier stages in which the transistors are in common base; the output impedance of the first stage and the input impedance of the second stage are too far apart.



h_o
(kΩ)
$-V_{CE}=2V$
$-I_C=2mA$
600
500
400
300
200
100
A
B
C
D
10
100
1000
10000
$R_g(\Omega)$

Fig. 426

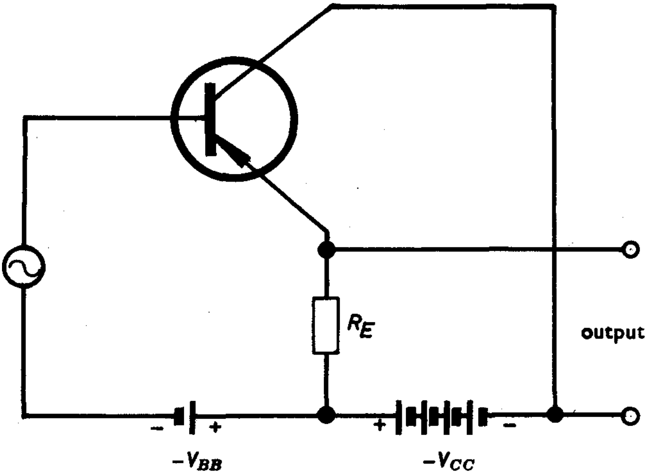

R_E
output
–
+
$-V_{BB}$
+
–
$-V_{CC}$

Fig. 427

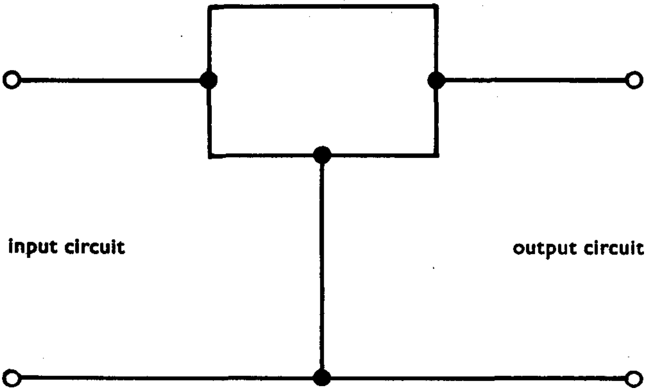

input circuit
output circuit

Fig. 428

CHAPTER 37

The common-collector configuration

In this case the generator which drives the transistor is connected between the base and the collector, while the load impedance across which the output signal appears is connected in series with the emitter (see Fig. 427). This corresponds to the equivalent circuit of Fig. 428, in which the input signal is applied between the base and the collector, while the output signal is taken off across the emitter and collector. The input and output circuits thus have the collector in common.

As in the previous circuits we shall investigate the influence of the load impedance on the power gain, and also the influence of the load impedance and the generator impedance on the input and output impedances of the transistor.

37.1. Power gain

In this case too the power gain is equal to the product of the voltage gain and the current gain. The voltage gain is once more equal to the ratio of the output voltage to the input voltage, while the current gain is given by the ratio of the output current to the input current.

Voltage gain

The voltage gain is equal to the quotient of the voltage produced across the resistance in series with the emitter (v_{R_E}) and the generator voltage v_g:

$$G_V = v_o/v_i = v_{R_E}/v_g.$$

The voltage gain of a transistor in common collector will always be very close to 1 and depends on the value of the resistance R_E in the emitter circuit. In Fig. 429, the relationship between R_E and G_V is plotted for constant values of $-V_{CE}$ and $-I_C$.

At $R_E = 1\ \mathrm{k}\Omega$ we have $G_V = 0.97$ (point A), at $R_E = 10\ \mathrm{k}\Omega$ the corresponding value is $G_V = 0.99$ (point B), at $R_E = 100\ \mathrm{k}\Omega$ the value of G_V is approximately 1 (point C) as also at $R_E = 1\ \mathrm{M}\Omega$ (point D).

The voltage gain thus increases extremely slowly with the value of the load impedance in the emitter circuit; at high values of R_E the voltage gain is practically constant and approximately equal to 1.

Current gain

The current gain is given by the quotient of the output current and the input current. From the equivalent circuit of Fig. 430 we see that the output current is equal to the emitter current, while the input current corresponds to the base current.

Now the emitter current is equal to the sum of the collector current and the base current, and since the latter may be neglected in relationship to the collector current, the current gain of a transistor in common collector may be represented by:

$$G_A = i_e/i_b.$$

This current gain is slightly greater than the value which can be obtained with a transistor in common emitter. ($G_A = i_c/i_b$). In contrast to a transistor in common base, a transistor in common collector does not give any voltage gain, but a very considerable current gain. (Practically equal to the current gain obtained in the common emitter configuration).

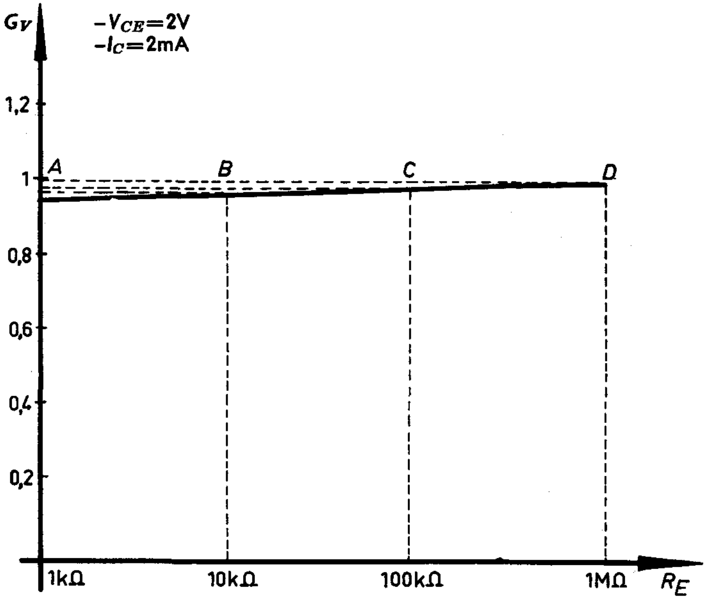


Fig. 429

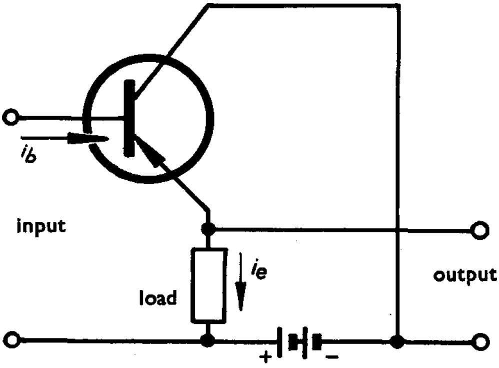

i_b
input
i_e
load
output
+
–

Fig. 430

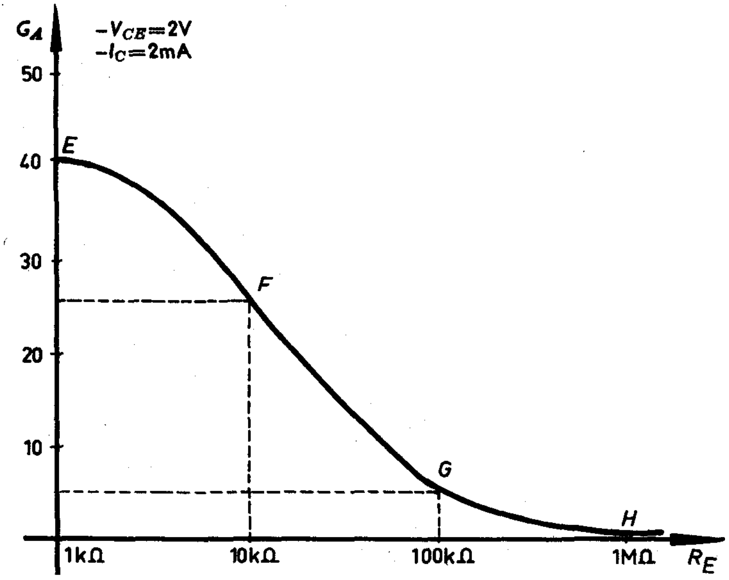


Fig. 431

The current gain of a transistor in common collector depends on the value of the resistance R_E which is included in the emitter circuit. The current gain G_A is plotted as a function of R_E in Fig. 431. At $R_E = 1\ \text{k}\Omega$ the current gain is 41 (point *E*), at $R_E = 10\ \text{k}\Omega$ it is 26 (point *F*) at $R_E = 100\ \text{k}\Omega$ the value is still 5.4 (point *G*) and at $R_E = 1\ \text{M}\Omega$ the current gain has dropped to about 1 (point *H*). The current gain of a transistor in common collector thus decreases rapidly as the load impedance increases, and particularly in the region between $R_E = 1\ \text{k}\Omega$ and $R_E = 100\ \text{k}\Omega$. The curve representing the current gain of a transistor in common collector as a function of the load impedance is practically identical with the one relating to the common emitter configuration.

Conclusion

In this case too the power gain is given by the product of the voltage gain and the current gain. Once more, we can determine the influence of the load impedance on the power gain with the aid of the curves which we have just discussed. Since the current gain of a transistor in common collector is practically equal to that obtained in common emitter, but the voltage gain is always much less, the power gain of a transistor in common collector will always be much less than the value obtained with a transistor in common emitter.

In order to determine the power gain of a transistor in common collector, the curves $G_V = f(R_E)$ and $G_A = f(R_E)$ are given once more in Figs. 432 and 433. The resulting power gain G_P is plotted in Fig. 434 as a function of the load impedance R_E. At $R_E = 1$ kΩ we have $G_V = 0.97$ (point A) and $G_A = 41$ (point E), from which it follows that the power gain is:

$$G_P = G_V G_A = 0.97 \times 41 = 40 = 16.2 \text{ dB (point } J).$$

At $R_E = 10$ kΩ the corresponding values are $G_V = 0.99$ (point B) and $G_A = 26$ (point E) from which the power gain follows as:

$$G_P = G_V G_A = 0.99 \times 26 = 26 = 14.5 \text{ dB (point } K).$$

At $R_E = 100$ kΩ we have $G_V = 1$ (point C) and $G_A = 5.4$ (point G), from which follows:

$$G_P = G_V G_A = 1 \times 5.4 = 5.4 = 7.4 \text{ dB (point } L).$$

At $R_E = 1$ MΩ the corresponding values are $G_V = 1$ (point B) and $G_A = 1.0$ (point H), from which the power gain is:

$$G_P = G_V G_A = 1 \times 1.0 = 1.0 = 0 \text{ dB (point } M).$$

The curve of Fig. 434 indicates that the maximum power gain is obtained with a load impedance less than 1 kΩ. This value of R_E corresponds to the output impedance of the transistor in common collector. At high values of R_E the power gain decreases rapidly with increasing resistance, but at lower values of R_E this quantity only has a slight effect on the power gain.

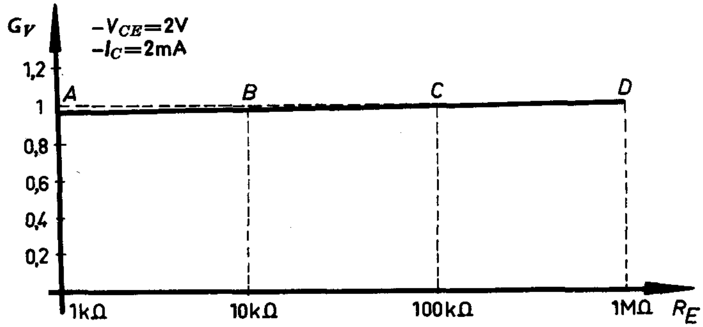


Fig. 432

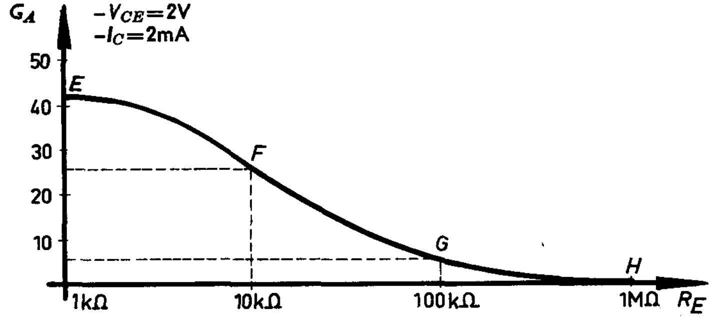


Fig. 433

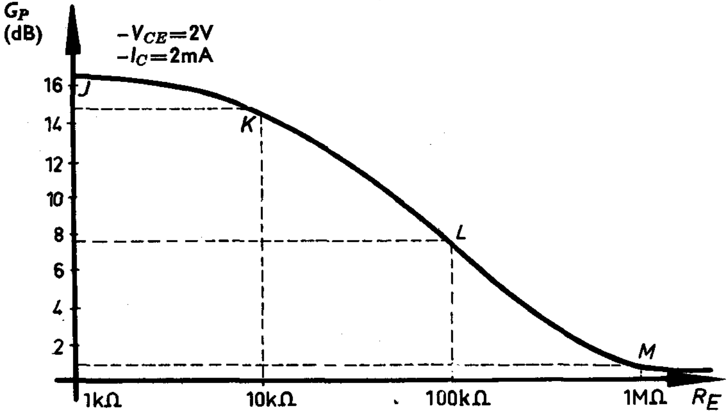


Fig. 434

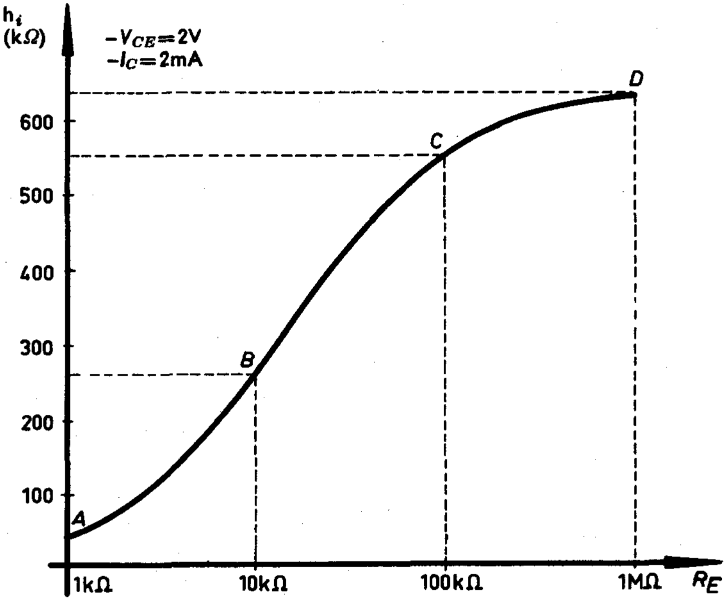

h_i
(kΩ)
$-V_{CE}=2V$
$-I_C=2mA$
600
500
400
300
200
100
A
B
C
D
1kΩ
10kΩ
100kΩ
1MΩ
R_E

Fig. 435

37.2. The influence of the load impedance on the input impedance of the transistor

The input impedance of a transistor in common collector does not only decrease as the collector current increases, but also depends greatly on the load impedance of the emitter circuit. The influence of this impedance R_E on the input impedance h_i is plotted graphically in Fig. 435. This curve shows that the input impedance increases with the load impedance.

At $R_E = 1$ kΩ the input impedance is $h_i = 42$ kΩ (point *A*),
at $R_E = 10$ kΩ the input impedance is $h_i = 260$ kΩ (point *B*),
at $R_E = 100$ kΩ the input impedance is $h_i = 550$ kΩ (point *C*),
at $R_E = 1$ MΩ the value is $h_i = 640$ kΩ (point *D*).

The input impedance of a transistor in common collector thus increases rapidly as the load impedance increases from 1 kΩ to 100 kΩ; at higher values of this impedance its influence on the input impedance of the transistor is not so great. The curve of (Fig. 436) was plotted at a constant value of $-V_{CE}$ and of $-I_C$; at a higher value of $-I_C$ the curve shifts in the direction of the abscissa.

In practice the input impedance of a transistor in common collector will be of the order of several hundred kilohms; this is the only configuration in which the input impedance is very much greater than the output impedance. This characteristic of the common-emitter configuration offers the possibility of numerous applications in radio receivers because the damping which the transistor exercises on the previous circuit is very slight.

37.3. The influence of the generator impedance on the output impedance of the transistor

The output impedance of a transistor in common collector decreases with increasing collector current but also depends on the generator impedance. In Fig. 436 the output impedance h_o is plotted as a function of the generator impedance R_g for $-V_{CE} = 2$ V and $-I_C = 2$ mA. From this curve we see that the output impedance of the transistor increases with the value of the generator impedance.

At $R_g = 10\ \Omega$ the output impedance is $h_o = 18\ \Omega$ (point *A*),
at $R_g = 100\ \Omega$ the output impedance is $h_o = 20\ \Omega$ (point *B*),
at $R_g = 1\ \mathrm{k}\Omega$ the output impedance is $h_o = 40\ \Omega$ (point *C*), and
at $R_g = 10\ \mathrm{k}\Omega$ the value is $h_o = 240\ \Omega$ (point *D*).

At values of the generator impedance lying between 10 Ω and 1 kΩ the output impedance only increases very slowly, but in the range between $R_g = 1\ \mathrm{k}\Omega$ and $R_g = 10\ \mathrm{k}\Omega$ the increase is much more rapid.

At higher values of the collector current we obtain similar curves, but these are situated nearer to the abscissa. In practice the output impedance of a transistor in grounded collector will be some tens of ohms, i.e. very much lower than the input impedance.

In the above considerations we have discussed the behaviour of the common collector configuration for the case in which the input signal is applied between the base and the collector, and in which the base thus forms the input electrode. In certain cases however it is also possible to apply the input signal between the emitter and the collector, in which case the emitter forms the input electrode. In this configuration the input impedance and the output impedance applicable when the base is the input electrode become interchanged, but the current gain is practically the same in both cases.

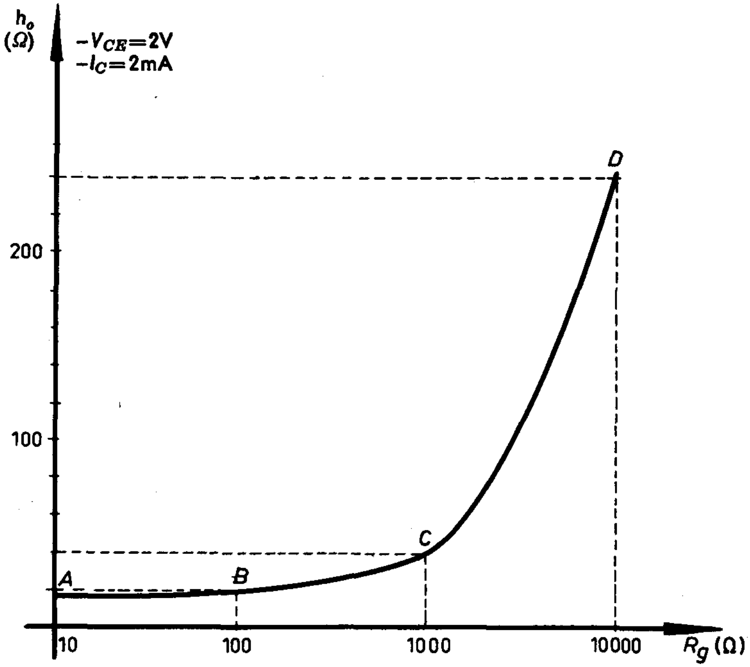


Fig. 436

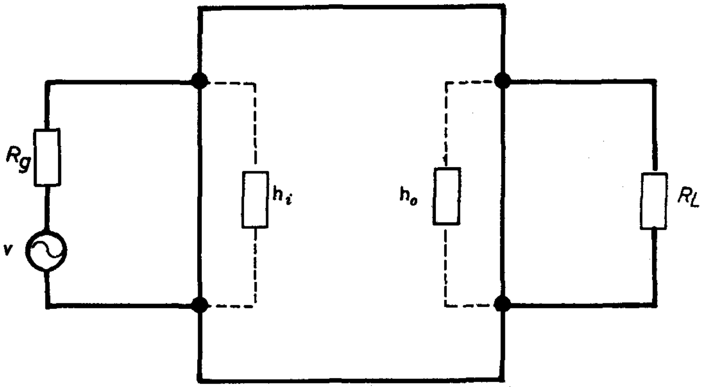


Fig. 437

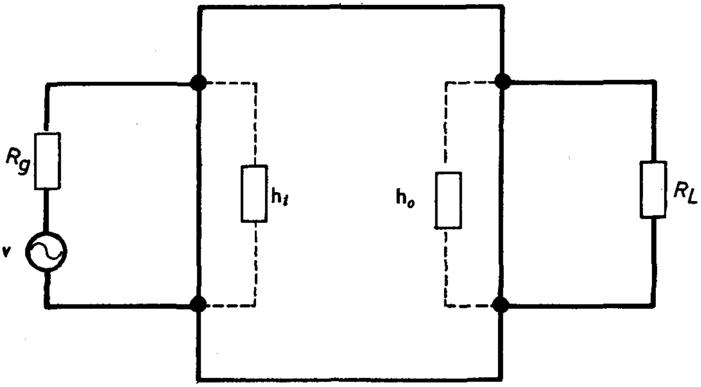


Fig. 438

CHAPTER 38

Comparison of the various configurations

We will now compare the various possible configurations by plotting the curves which we have considered in the previous chapters in one and the same graph. These curves are characteristic for the operation of the transistor and from them we can draw conclusions with regard to the suitability of any given circuit.

To this end we shall compare the following five characteristics for the three configurations in succession:

The voltage gain and the way in which it is influenced by the load impedance.

The current gain and the way in which it is influenced by the load impedance.

The power gain and the way in which it is influenced by the load impedance.

The input impedance and the manner in which it is influenced by the load impedance, (that is, the collector load in common emitter and common base and the emitter load in the common collector configuration, see Fig. 437).

The output impedance and the way in which it is affected by the generator impedance (see Fig. 438).

In most cases the principal requirement for an amplifier stage is that it should give the greatest possible power gain. From the considerations of section 38.3 concerning the power gain it is possible to say which configuration will be the best for any given purpose.

The power gain that can be obtained also depends on the matching between two stages, and this is principally determined by the input and output impedances of the transistors which are used. It is thus very important to balance these various factors accurately against each other in order to determine how the maximum power gain can be obtained.

38.1. The voltage gain and its dependence on the load impedance

In the common emitter and common base configurations the voltage gain reaches a high value; in the common collector configuration, on the other hand, the voltage gain is always in the region of 1. In Fig. 439 the voltage gain for the various configurations is plotted as a function of the load impedance R_L. The black curve relates to the common emitter configuration, the green curve to common base, and the red curve to common collector. For common emitter and common base we see that the voltage gain is practically the same, whatever the value of R_L. In both cases the voltage gain increases with R_L.

By contrast, the voltage gain in the common collector configuration is very small and is practically equal to 1; the output voltage follows the input voltage and is always slightly less than the latter, as in the case of a thermionic valve connected as a cathode follower. It is this common collector configuration which ensures the most constant voltage gain if the load impedance connected in series with the emitter is subject to variation. In every case the voltage gain increases with the value of the load impedance, but the latter cannot be increased indefinitely because of the limits set for the supply voltage and the constant collector current.

38.5. The output impedance and its dependence on the generator impedance

The output impedance of a transistor in common emitter is several tens of kilohms, in common base several hundreds of kilohms or even about 1 megohm, and in common collector several tens of ohms, (when the base is used as the input electrode).

In all three configurations the output impedance is dependent on the generator impedance, as we see from Fig. 443 in which this output impedance is plotted against the generator impedance. The black curve relates to the common emitter configuration, the green curve to common base and the red curve to common collector.

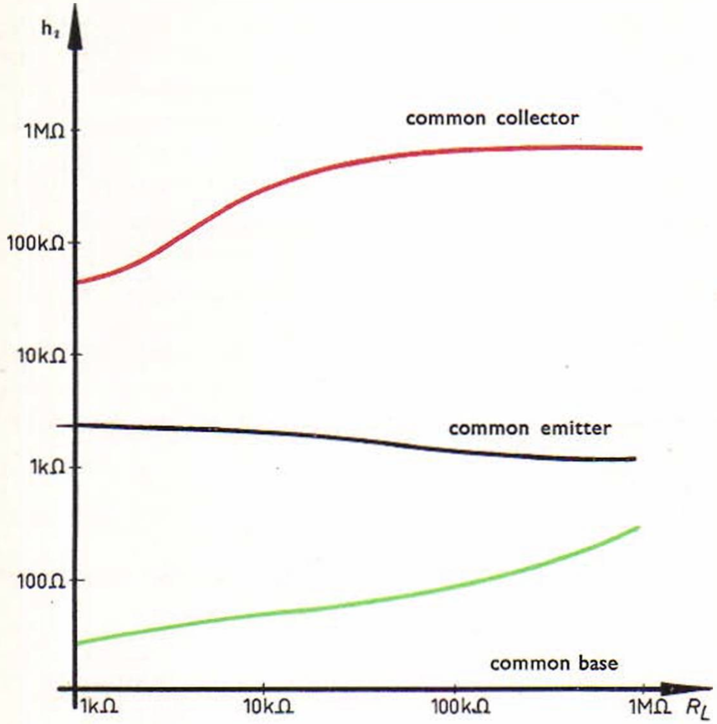
In common emitter the output impedance of the transistor decreases slightly with increasing generator impedance, while in the other two configurations the output impedance increases with the generator impedance. In common emitter the influence of the generator impedance on the transistor output impedance is relatively slight; in common collector however, this influence is fairly large, particularly at relatively high values of the generator impedance. For example, the output impedance at a generator impedance of 10 Ω is 18 Ω, and for a generator impedance of 10 kΩ it is 360 Ω, so that a change of 1 : 1,000 in the generator impedance results in a change of 1 : 20 in the output impedance. In common base a change of this size in the generator impedance results in the output impedance increasing in the ratio 1 : 12.

38.4. The input impedance and its dependence on the load impedance

The input impedance of a transistor in common emitter is of the order of 1 kilohm, in common base it is of the order of some tens of ohms and in common collector of the order of some hundreds of kilohms.

As already explained, the input impedance depends on the load impedance. In Fig. 442 the input impedance is plotted as a function of the load impedance, for all three configurations. The black curve relates to the common emitter configuration, the green curve to common base and the red curve to common collector.

In common emitter the input impedance decreases as the load impedance increases, while in both other configurations the reverse is true. This effect is explained by the fact that in the common base and common collector configurations the input and output signals are in phase, as a result of which internal feedback occurs. In common emitter on the other hand, the input and output signals are in phase opposition so that there is no question here of internal feedback. Fig. 442 also shows that the common collector configuration is the one in which the input impedance depends most on the load impedance.



h_I
1MΩ
100kΩ
10kΩ
1kΩ
100Ω
common collector
common emitter
common base
1kΩ
10kΩ
100kΩ
1MΩ
R_L

Fig. 442

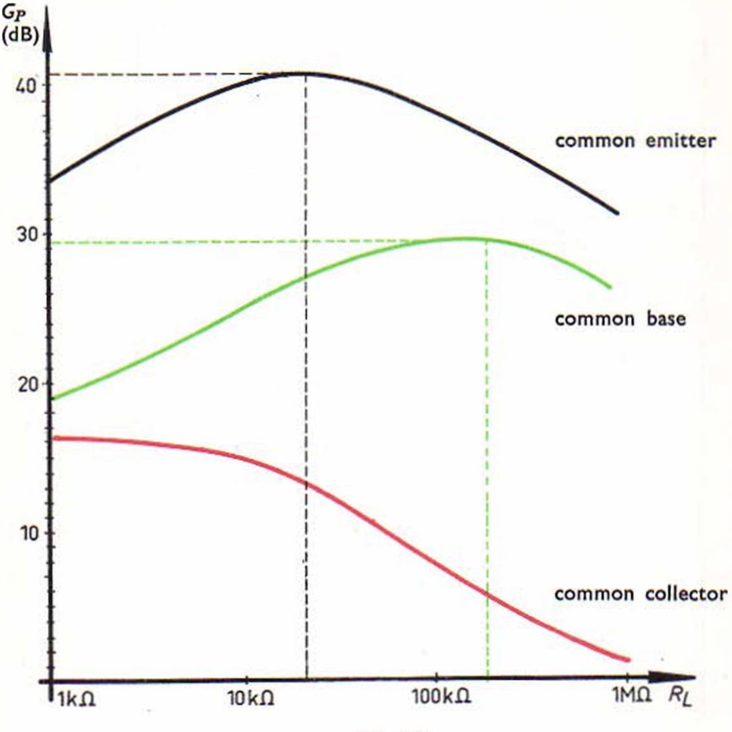

G_P (dB)
40
30
20
10
common emitter
common base
common collector
1kΩ
10kΩ
100kΩ
1MΩ
R_L

Fig. 441

38.3. The power gain and its dependence on the load impedance

In the common emitter configuration both the voltage gain and the current gain reach high values, so that the power gain is also high. In common base the voltage gain is practically equal to the voltage gain in common emitter, but the current gain is less than one, so that the power gain in this case is much smaller. In the common collector configuration the voltage gain is approximately equal to one; it is true that the current gain is slightly higher than the gain obtained in common emitter, but in practice the power gain will always be less. In Fig. 441 the power gain in the three configurations is plotted as a function of the load impedance R_L.

Once again the black curve relates to the common emitter configuration, the green curve to common base and the red curve to common collector.

This graph shows that the maximum power gain is obtained under the following conditions:

In common emitter, when the load impedance has a value of approximately 20 kΩ.
In common base, when the load impedance has a value of approximately 200 kΩ.
In common collector, when the load impedance in the emitter circuit is less than 1 kΩ.

In all three circuits the power gain is a maximum when the load impedance equals the output impedance of the transistor. The latter is thus

20 kΩ in common emitter,
200 kΩ in common base, and
some tens of ohms in common collector.

The power gain in common emitter is higher than the values which can be obtained in the other configurations whatever the value of the load impedance.

38.2. The current gain and its dependence on the load impedance

In the common emitter and common collector configurations the current gain is considerable; in common base on the other hand, it is always less than 1. In Fig. 440 the current gain is plotted as a function of the load impedance R_L for the three configurations. The black curve relates to the common emitter configuration, the green curve to common base and the red curve to common collector.

The graph shows that the current gain and its dependence on the load impedance is practically the same for common emitter and common collector. In common base however, the current gain is always less than 1 and the output current follows the value of the input current. In every case the current gain decreases as the load impedance increases and the maximum current gain is obtained at the minimum load impedance. It should be remembered that the maximum current gain does not correspond to the maximum power gain; this is because the former is obtained when the load impedance equals zero and in that case the voltage gain, and therefore the power gain as well, also equals zero.

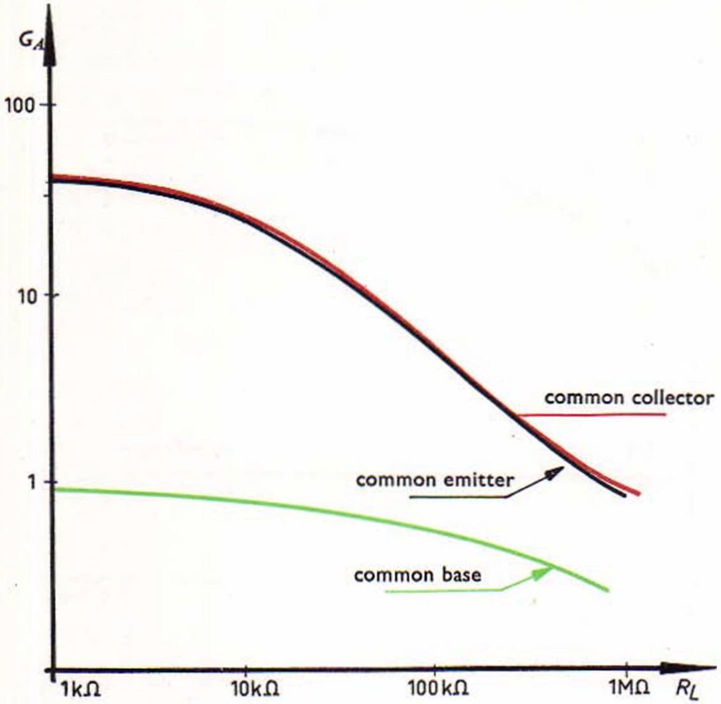

G_A
100
10
1
common collector
common emitter
common base
1kΩ
10kΩ
100kΩ
1MΩ
R_L

Fig. 440

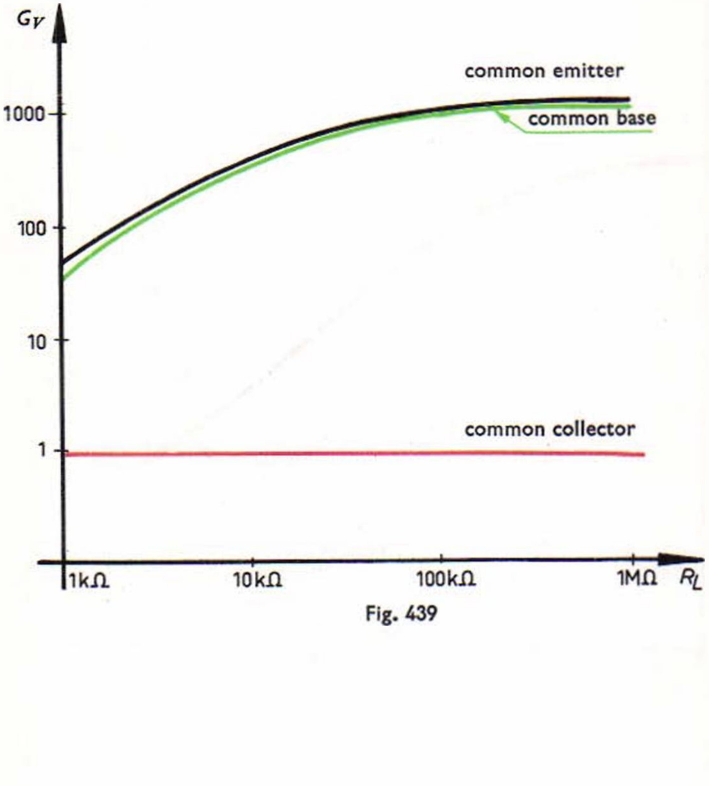

G_V
1000
100
10
1
common emitter
common base
common collector
1kΩ
10kΩ
100kΩ
1MΩ
R_L

Fig. 439

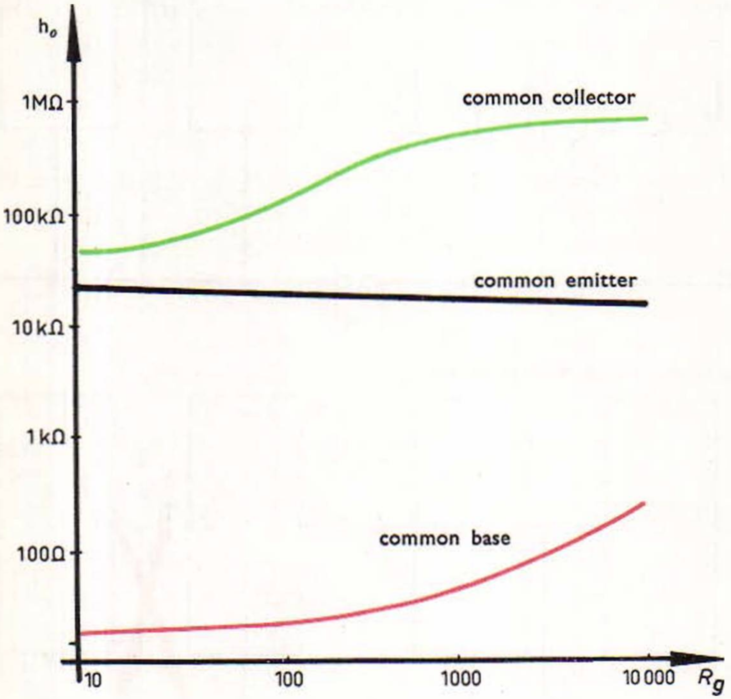

h$_o$
1MΩ
100kΩ
10kΩ
1kΩ
100Ω
common collector
common emitter
common base
10
100
1000
10 000
R_g

Fig. 443

common emitter circuits in cascade

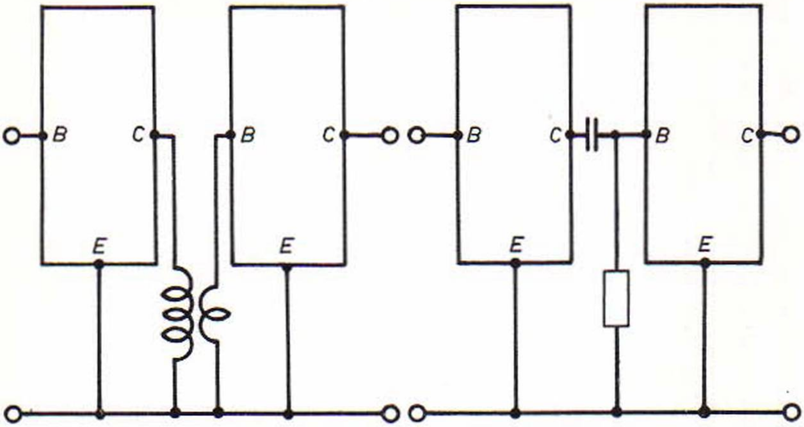

B
C
B
C
B
C
B
C
E
E
E
E

Fig. 444

common base circuits in cascade

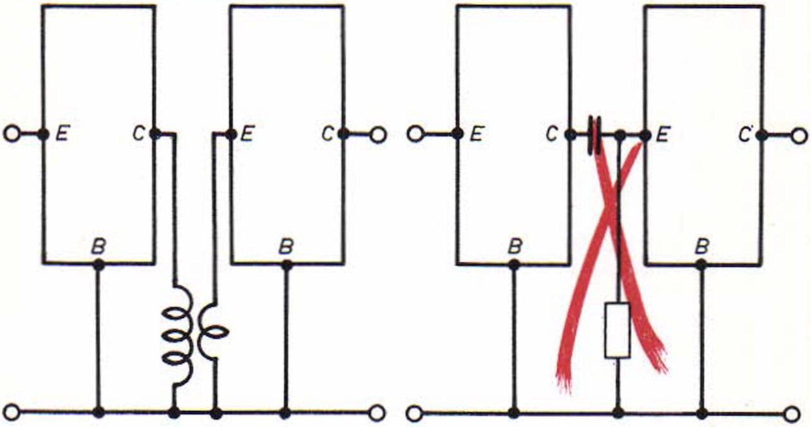

E
C
E
C
E
C
E
C
B
B
B
B

Fig. 445

38.6. Conclusions

The common-emitter configuration offers two advantages over the other two configurations: the power gain which can be obtained is considerably greater, and the difference between the input impedance and the output impedance of the transistor is relatively small ($h_i \simeq 1\ \mathrm{k}\Omega$, $h_o \simeq 10\ \mathrm{k}\Omega$ to $20\ \mathrm{k}\Omega$).

In an amplifier consisting of a number of transistors in common emitter, connected in cascade, the coupling element may either be a transformer (in order to obtain optimum matching) or an *CR*-coupling (Fig. 444).

The power gain obtainable in the common-base configuration is appreciably less; in addition the difference between the input impedance and the output impedance of the transistor is very great ($h_i \simeq 50\ \Omega$, $h_o \simeq 200\ \mathrm{k}\Omega$). If it is required to use a number of transistors in common base in an amplifier, these cannot be coupled by means of a resistance and a capacitance. The efficiency of such an arrangement would be extremely low, so that only transformer coupling could be employed (Fig. 445).

The common-collector configuration gives a still lower power gain. The input impedance in this case is very high, while the output impedance is low (h_i of the order of several hundred kilohms, h_o of the order of some tens of ohms). However, this configuration offers the advantage of an extremely high input impedance.

In certain applications in which the use of the common emitter configuration would result in too much damping of the previous circuit, it can be recommended that a transistor in common collector should be used as an intermediate stage, in order to obtain good impedance matching.

Appendix

In addition to the characteristics we have discussed, the transistor manufacturer also supplies graphs showing the influence of the collector current and voltage on the various h parameters.

These parameters correspond to the elements which determine the electrical behaviour of the transistor under certain conditions and for small signals. Our remarks on the families of curves for $-I_C = f(-V_{CE})$, $-I_C = f(-I_B)$, $-I_B = f(-V_{BE})$ and $-V_{BE} = f(-V_{CE})$, in connection with which we discussed the various electrical characteristics of the transistor, do in fact refer exclusively to signals of small amplitude.

We can also draw a parallel between the various internal resistances of the transistor, the current gain, the internal feedback and the parameters h_{ie}, h_{oe}, h_{fe}, and h_{re}. The parameters follow from the families of curves given in Fig. 446. The connection with the h parameters is as follows:

h_{ie} = input impedance of the transistor with output short – circuited: this follows from the $-I_B = f(-V_{BE})$ characteristic.

$1/h_{oe}$ = output impedance of the transistor with input short-circuited: this follows from the $-I_C = f(-V_{CE})$ characteristic.

h_{re} = internal feedback of the transistor with input open-circuited: this follows from the $-V_{BE} = f(-V_{CE})$ characteristics.

Consequently it is possible to deduce the parameters as functions of the collector current and of the collector-emitter voltage from the complete families of curves. It is in fact simpler, however, to base ones considerations on the characteristics and their variations, which are mentioned in the transistor data.

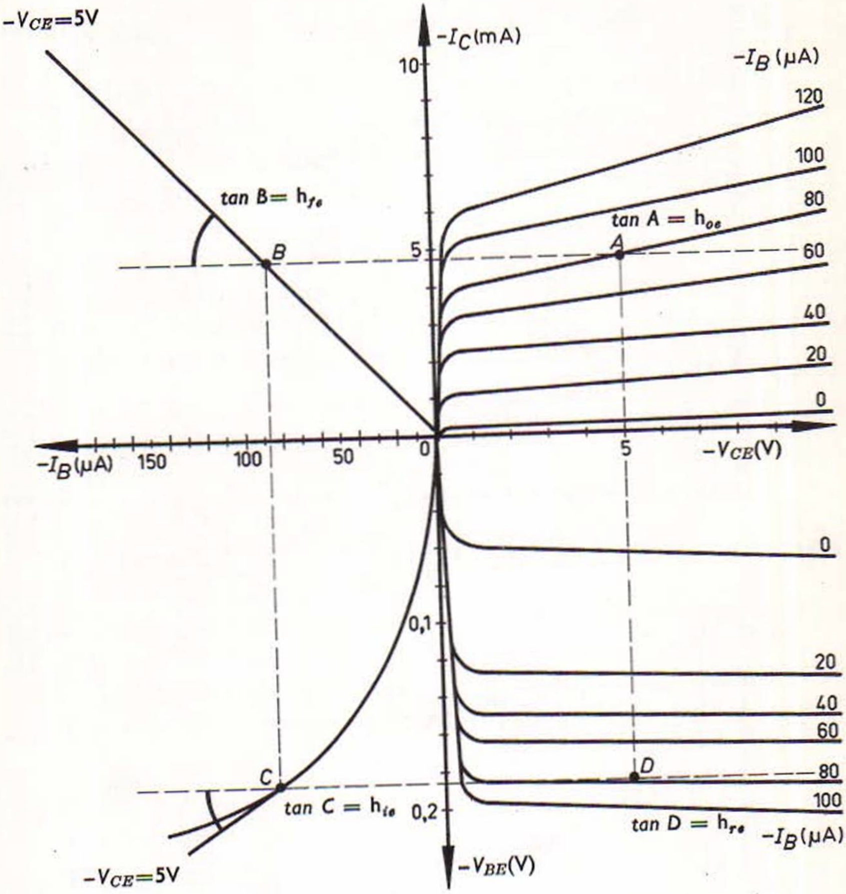

-V_CE=5V
-I_C(mA)
10
-I_B (µA)
120
100
tan B= h_fe
80
tan A = h_oe
B
5
A
60
40
20
0
-I_B(µA)
150
100
50
0
5
-V_CE(V)
0
0,1
20
40
60
C
D
80
tan C = h_ie
0,2
100
tan D = h_re
-I_B(µA)
-V_CE=5V
-V_BE(V)

Fig. 446

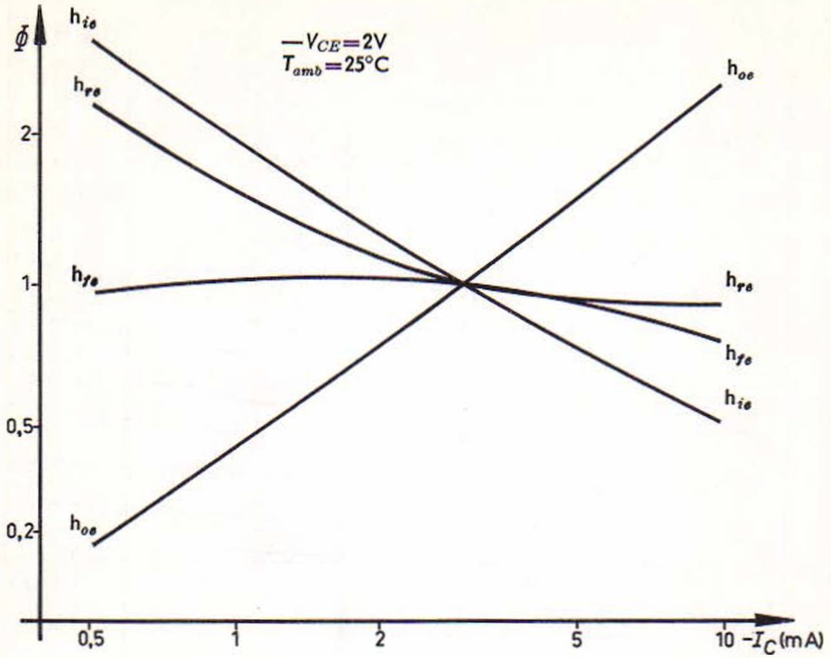


Fig. 447

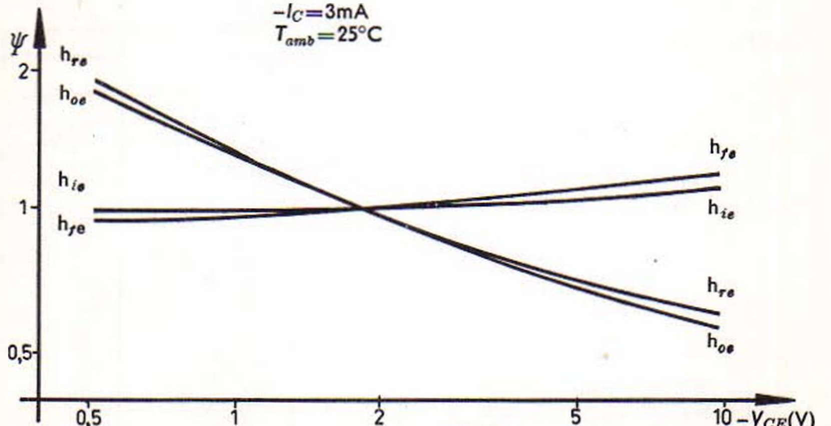


Fig. 448

The influence of the collector current on the h parameters

In order to be able to represent the influence of the collector current on the four h parameters in one graph, we mark out the ordinate in values of the quantity Φ i.e. the ratio of the value of the given parameter for an arbitary collector current to its value for a given collector current $-I_C$, (In Fig. 447 the value of this current is 3 mA). In this way, it is possible to use a single standard scale for the ordinate. The manufacturer gives the value of the various h parameters for $-I_C = 3$ mA and $-V_{CE} = 2$ V.

For the OC 71 transistor for example, the values at $-V_{CE} = 2$ V, $-I_C =$ 3 mA and $T_{\text{amb}} = 25°$ C are:

h_{ie} (input impedance) = 800 Ω.
$1/h_{oe}$ (output impedance) = 12.5 kΩ.
h_{fe} (current gain) = 47.
h_{re} (internal feedback) = 5.4×10^{-4}.

If we now wish to know the value of one of these parameters for a collector current other than 3 mA all that is necessary is to consult the corresponding curve in Fig. 447 and to multiply the above value by the ratio Φ for the required value of $-I_C$. The following example will illustrate this.

Suppose that we wish to find the value of h_{ie} (input impedance with short-circuited output) for an OC 71 transistor at a collector-emitter voltage of 2 V and a collector current of 1 mA. As given above, the value of h_{ie} for $-I_C = 3$ mA is 800 Ω. At a collector current of 3 mA the curve h_{ie} corresponds to the value 1 on the ordinate, and for a collector current of 1 mA the curve gives the value 2 on the ordinate, so that $\Phi = 2/1 = 2$. This means that for a collector current $-I_C = 1$ mA, the parameter $h_{ie} = 2 \times 800 = 1600\ \Omega$.

The influence of the collector-emitter voltage on the h parameters

A similar graph representing the influence of the collector-emitter voltage on the h parameters is given in Fig. 448. Here too the ordinate is calibrated in terms of a standard quantity Ψ which represents the ratio between the value of the parameter in question for an arbitary collector-emitter voltage and its value for a given collector-emitter voltage ($-V_{CE} = 2$ V). This graph relates to $-I_C = 3$ mA. and, as in the previous case, it can be used to find the value of any h parameter for any collector-emitter voltage from the known value of this parameter at $-V_{CE} = 2$ V.

DIODES AND TRANSISTORS FONTAINE

PHILIPS + TECHNICAL + LIBRARY